WIRELESS ENGINEER

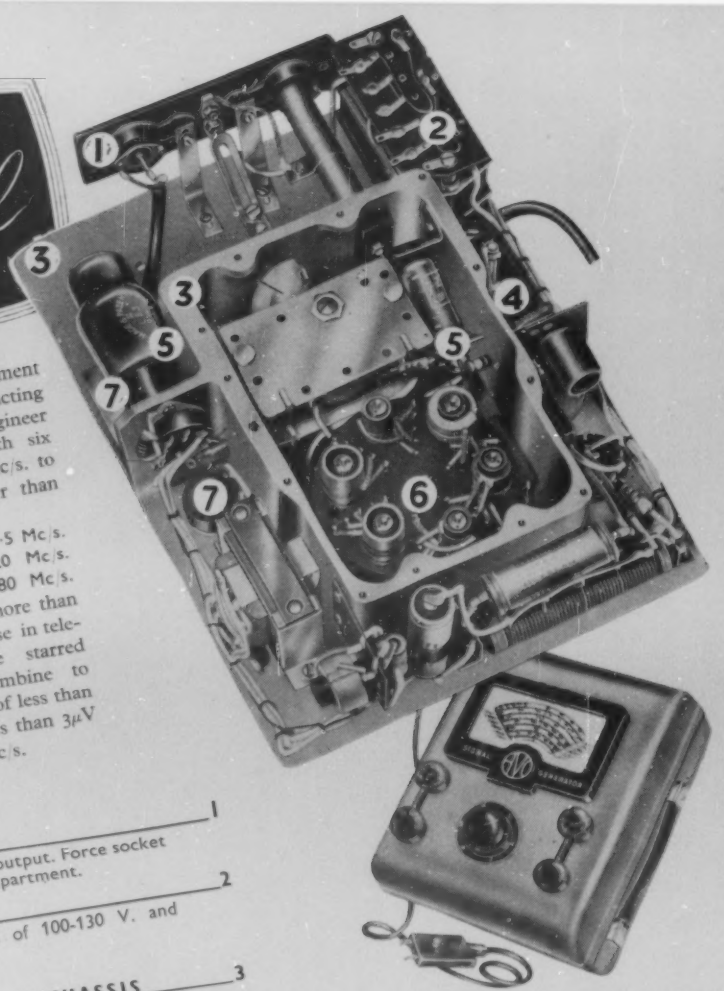
THE JOURNAL OF RADIO RESEARCH & PROGRESS

FEBRUARY 1954

VOL. 31 No. 2 THREE SHILLINGS AND SIXPENCE

avo

WIDE RANGE *Signal* GENERATOR



An inexpensive yet precision instrument designed specially to meet the exacting needs of the modern service engineer and laboratory technician. With six frequency ranges covering 50 Kc/s. to 80 Mc/s., its accuracy is better than ± 1% of the scale reading.

50 Kc/s.-150 Kc/s.	1·5 Mc/s.-5·5 Mc/s.
150 Kc/s.-500 Kc/s.	5·5 Mc/s.-20 Mc/s.
500 Kc/s.-1·5 Mc/s.	20 Mc/s.-80 Mc/s.

Scale sub-divisions provide more than adequate discrimination for use in television circuits. Note the starred features below, which combine to maintain a minimum signal of less than 1μV up to 20 Mc/s. and less than 3μV between 20 Mc/s. and 80Mc/s.

1 ★**OUTPUT**
Co-axial socket for attenuated output. Force socket located totally within H.F. compartment.

2 **MAINS TRANSFORMER**
Marked tagboard for inputs of 100-130 V. and 200-260V., A.C. 50/60 c/s.

3 ★**CAST ALUMINIUM H.F. COMPARTMENT and CHASSIS**
Large number of fixing holes for H.F. compartment cover ensures excellent electrical bonding and good screening.

4 **FUSE**
Easily accessible when replacement is necessary.

5 **VALVES**
Standard types run at a rating to ensure long life.

6 **TURRET COIL SWITCHING**
Standard "AVO" practice.

7 ★**ATTENUATOR SYSTEM**
Employs close tolerance, high stability midget carbon resistors, low reactance rotary potentiometer modified for H.F. operation with carefully designed screening.

Other features include:

ILLUMINATED SPOT RANGE SELECTOR
Gives rapid identification of operational band with intensified lighting round precise frequency. Fine hair line gives close discrimination, particularly on high frequencies.

STOPS
Separate stops prevent turning of dial with respect to condenser.

★**MAINS FILTER SYSTEM**
This is screened from main electrical assembly.

BUSHING PLATES
provide additional rigidity for rotary controls.

SLOW MOTION DRIVE
Substantially free from backlash.

MAINS MODEL 100—130 V. and 200—260 V. A.C., 50—60 c/s. OR BATTERY MODEL..... £30.

Sole Proprietors and Manufacturers:—

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WINDER HOUSE • DOUGLAS STREET • LONDON • S.W.1 *Telephone: VICTORIA 3404/9*

S.G.2

'VARIAC' voltage regulating transformers

Reg'd Trade Mark

With a 'VARIAC' voltages are instantly and minutely adjustable from 0-Line Voltage, or in some cases up to 17% above line voltage. Type 50-B 'VARIAC,' as illustrated left, is often operated in a 3-gang assembly on 3-phase work to control 21kVA.

Type 200 C.U.H. 'VARIAC'

Type 100-R 'VARIAC'

SERIES 50 'VARIAC' TRANSFORMERS.

SPECIFICATIONS

Type	Load Rating	Input Voltage	Current		Output Voltage	No-Load Loss	Net Price £ s. d.*
			Rated	Maximum			
50-A	5 kva.	115 v.	40 a.	45 a.	0-135 v.	65 watts	44 18 6
50-B	7 kva.	230/115 v.	20 a.	31 a.	0-270 v.	90 watts	44 18 6

All 'VARIAC' prices plus 20% as from 23rd Feb. 1952

REQUEST ALSO OUR 20-PAGE SUPPLEMENTARY CATALOGUE GIVING COMPLETE INFORMATION ON OUR NEW AND COMPLETE RANGE OF AC AUTOMATIC VOLTAGE STABILISERS: THESE RANGE FROM 200 VA TO 25 kVA. PERFORMANCE IS EXCELLENT, FROM NO-LOAD TO FULL LOAD, AND STABILITY IS QUITE UNAFFECTED BY FREQUENCY VARIATIONS.

PRICES ARE EXCEPTIONALLY LOW.

Write for catalogue V549 which gives full details of 'VARIAC' transformers and suggestions for use.

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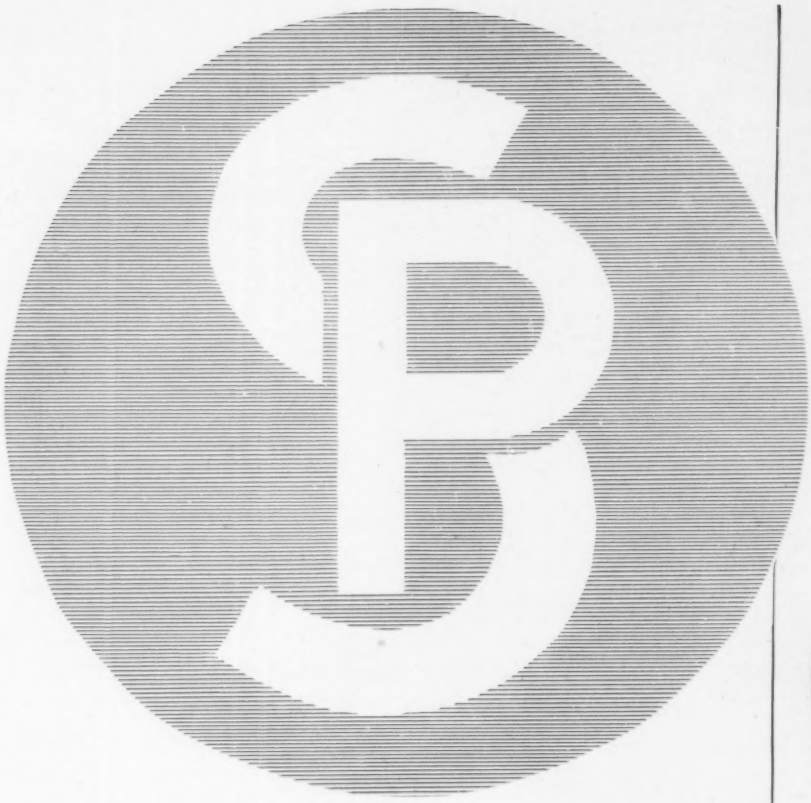

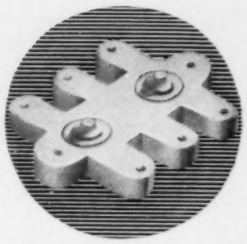

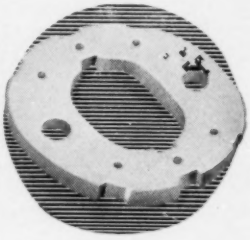

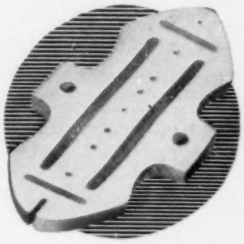

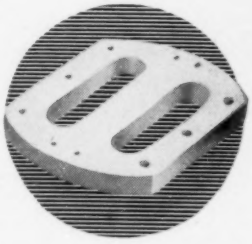

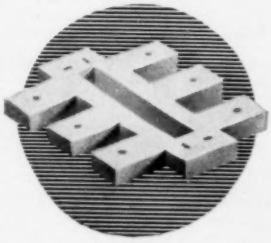

for
radio
ceramics

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SP87

Mullard Ferroxcube being extruded into rods for H.F. cores.

MAGNETIC MATERIALS Extensive research and manufacturing facilities have established Mullard as the leading producers of magnetic materials. They were the first, for example, to introduce Ferroxcube, the world's most efficient magnetic ferrite; 'Ticonal' anisotropic permanent magnets, renowned for their high stability and high energy output; and Magnadur, an entirely new type of permanent magnet with the insulating properties of a ceramic.

The wealth of experience gained from these developments is available to all users of magnetic materials through the Mullard advisory service. An enquiry to the address below will put a team of specialised engineers at your disposal.

Mullard **Mullard**

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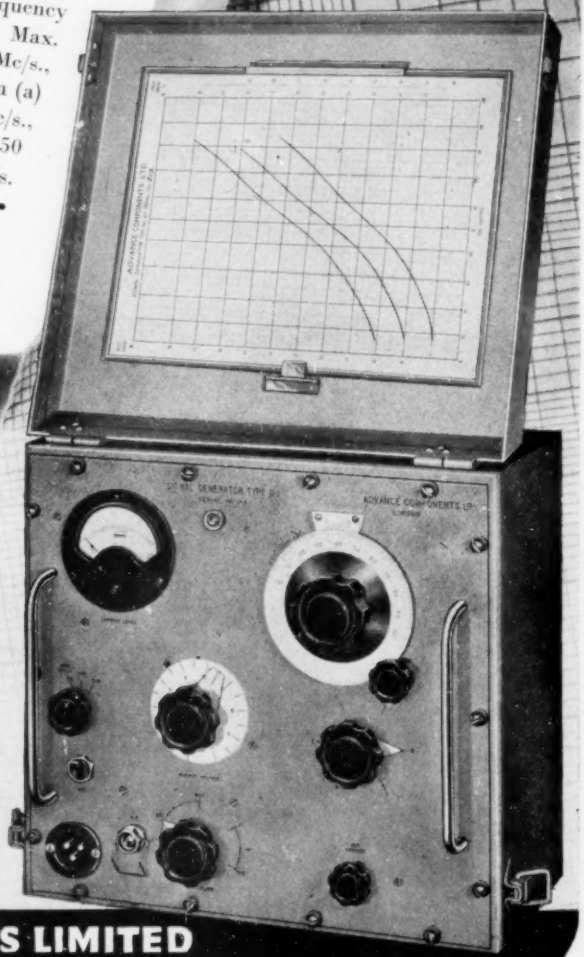
MULLARD LTD., COMPONENT DIVISION, CENTURY HOUSE, SHAFTESBURY AVENUE, LONDON, W.C.2.

from 10 to 310 mc/s

Within the small compass of $14\frac{1}{4}'' \times 12\frac{3}{4}'' \times 8''$, this generator provides facilities as are normally available only in instruments of much greater size and weight—and with an accuracy which suggests something far more costly. Outstanding features include: Frequency calibration ± 1% • Max. attenuation error at 300 Mc/s., ± 4 db. • Modulation (a) 30% sine wave at 1,000 c/s., (b) approximately 50/50 square wave at 1,000 c/s. • Negligible stray field • Weight 34 lb.

Fullest details contained in leaflet V6 sent on request.



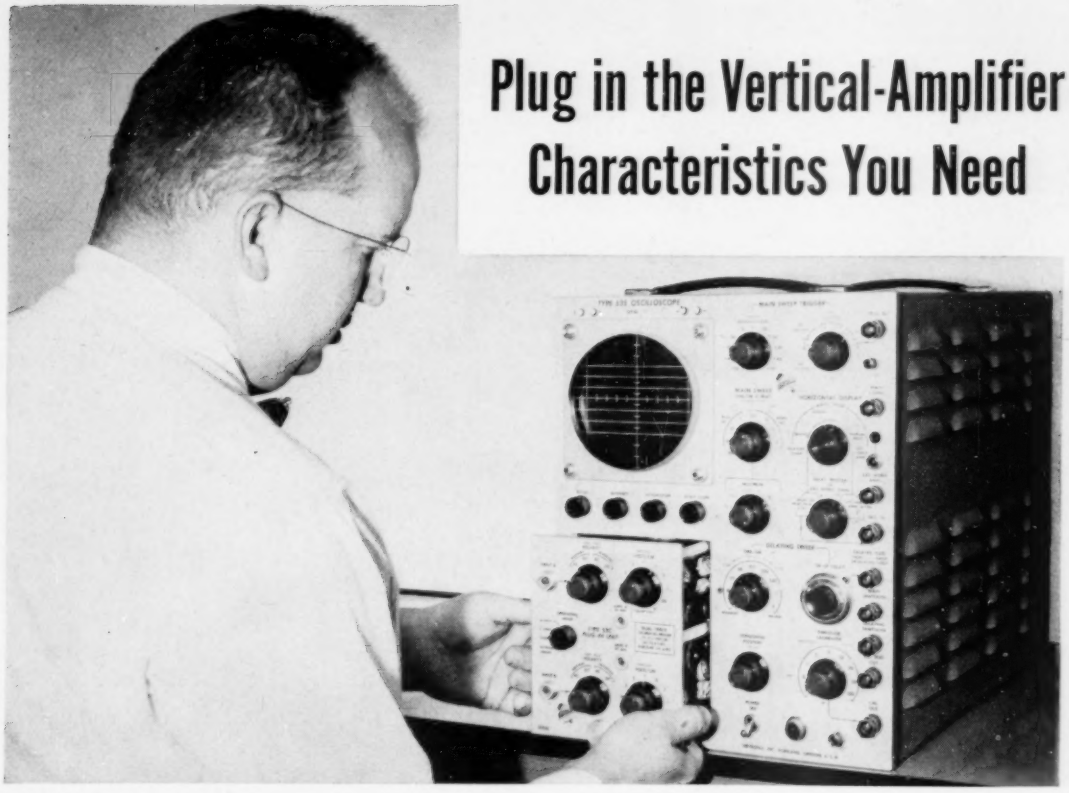
D1

Advance

V.H.F. SIGNAL GENERATOR

ADVANCE COMPONENTS LIMITED

BACK RD., SHERNHALL ST., WALTHAMSTOW. E.17. *Phone*: Larkswood 4366



Plug in the Vertical-Amplifier Characteristics You Need

...with this new Tektronix Oscilloscope

No need to switch oscilloscopes when your requirements change. An extremely wide range of work is now handled by one oscilloscope—the Tektronix Type 535—with its moderately-priced plug-in units.

WIDE-BAND PLUG-IN PREAMPLIFIER. Bandpass dc to 10 mc, risetime 0.035 μsec, calibrated sensitivity to 0.05 v/cm, two signal inputs with more than 60 db isolation.
Type 53A Wide-band Plug-in Unit—$85

HIGH-GAIN PLUG-IN PREAMPLIFIER. 1 millivolt/cm sensitivity dc to 250 kc with bandpass increasing to dc to 750 kc at 50 mv/cm sensitivity and lower, differential input.
Type 53D High-gain Plug-in Unit—$145

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The use of plug-in preamplifiers providing this wide range of facilities is made possible by the broad basic specifications of the Type 535.

600,000,000 TO 1 TIME-BASE RANGE
0.02 μsec/cm to 12 sec/cm.

10 KV ACCELERATING POTENTIAL
Brighter display at low repetition rates.

NEW FLEXIBLE SWEEP DELAY
Accurate, jitter-free, 5 μsec to 20,000 μsec.

Type 535 Oscilloscope—$1300 plus price of desired plug-in unit. Prices f.o.b. Portland, Oregon

Please write for complete specifications

Tektronix, Inc.

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TYPE D-650-A

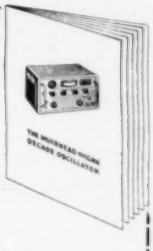
The *NEW* Muirhead-Wigan decade oscillator

THIS precision laboratory oscillator, which covers a range of 1 to 111,100c/s with an overall frequency accuracy of ±0.2% or ±0.5c/s, employs the decade tuning system, by means of which the frequency can be set quickly and accurately on four decade dials and a range switch. This system of tuning ensures the highest possible frequency accuracy and stability. It also enables a given frequency setting to be repeated exactly, and permits the addition or subtraction of a fixed number of cycles per second, thus giving an incremental accuracy of an extremely high order. No other type of oscillator possesses all these advantages.

FEATURES

Frequency range: 1-11,110c/s and 10-111,100c/s.
Frequency accuracy: ±0.2% or ±0.5c/s.
Hourly frequency stability: ±0.02% over most of range.
Maximum output:
2W into 8000 ohms above 20c/s.
50mW into 8000 ohms below 20c/s.
Harmonic content: 1% at 1W output.
Hum level: —80db relative to maximum output at 1000c/s.
Power supply: 200-250V, 50c/s; 90W. or 95-125V, 60c/s; 90W.
Dimensions: 17¼ in. wide x 10½ in. high x 13 in. deep. (43·8 cm x 26·7 cm x 33 cm).
Weight: 83 lb. (38 kg).

MAIL THIS COUPON FOR DESCRIPTIVE BROCHURE



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Close technical control of production is an all-important factor in the manufacture of Telcon Metals. Accurate tests and measurements are made at various stages in the course of production and the illustrations show two of the many routine examinations conducted in our well-equipped laboratories.

Precision measurement of magnetic characteristics of toroidal Mumetal cores by the Ferrometer.

Routine Factory testing of Mumetal toroidal cores.

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MAGNETIC ALLOYS — Mumetal, Radiometal, H.C.R., Rhometal, Permendur, R2799, 36/64, Dust.

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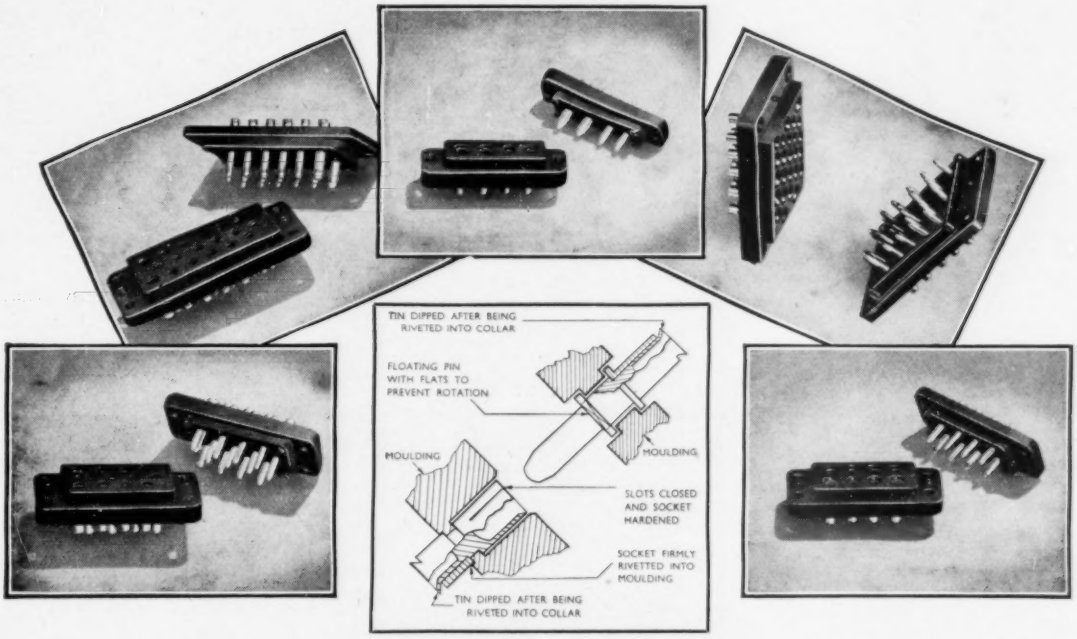
'UNITORS'

A range of miniaturised connectors by "Belling-Lee"


TIN DIPPED AFTER BEING RIVETED INTO COLLAR
FLOATING PIN WITH FLATS TO PREVENT ROTATION
MOULDING
MOULDING
SLOTS CLOSED AND SOCKET HARDENED
SOCKET FIRMLY RIVETTED INTO MOULDING
TIN DIPPED AFTER BEING RIVETED INTO COLLAR

4 to 25 WAY FROM STOCK

"Unitors" are links in the chain of electronic events. Any weakness means failure, and perhaps disaster. "Belling-Lee" Unitors are approved by A.I.D. and A.R.B., and have Joint Service Approval.

In recent independent tests, five pairs of 12-way "Unitors" made 16,335 insertions and withdrawals and were still serviceable.

The test apparatus was rigged in such a way that the plug portion was offered up to the socket with the pins at the maximum limit of float, and individual pins approached the relevant socket at various angles.

A coupling comprises a block of plugs and a complementary block of sockets, arranged so as to be non-reversible. Round pins are employed, turned from high grade brass, silver plated. The normal pins carry 3 amp., but each block has two large pins or sockets to carry 10 amp. All pins are fully floating, and the method of assembly is the subject of a patent.

Resilient sockets are of differentially hardened beryllium copper. All contacts, plugs and sockets are numbered on the face and reverse sides of the body. The distance between flanges when plugged together is 0·281 in.

U.K. Pat. No. 649739

Swiss Pat. No. 277799

A cable cover for line connection is available, together with retainer for chassis mounting.

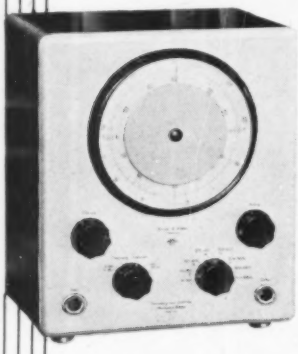
LIST NO.	PINS
L.653 P & S	4
L.654 P & S	8
L.655 P & S	12
L.656 P & S	18
L.657 P & S	25

BELLING & LEE LTD

GREAT CAMBRIDGE ROAD, ENFIELD, MIDDX., ENGLAND

FOR MEASURING FREQUENCY and DISTORTION

The model 1602 **Frequency and Distortion Measuring Bridge** is a simply operated blocking network, tuneable through the range from 47 to 12.000 c/s and attenuating the fundamental frequency 120 db.



type 1602
Tuneable 120 db attenuator

Wherever an interfering frequency must be blocked out, such as a hum voltage, the sharp attenuation provided by the instrument assures the complete elimination of the unwanted signal. Harmonic components may be measured in conjunction with the **Frequency Analyzer** model 2105. This versatile frequency-selective V.T. voltmeter, used over the whole field of acoustical and electro-acoustical measurements, gives in this case the harmonics as percentages of the amplitude of the fundamental note down to 0·01%.

type 2105
Multipurpose Audio Frequency Analyzer

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Full information will be supplied on request.

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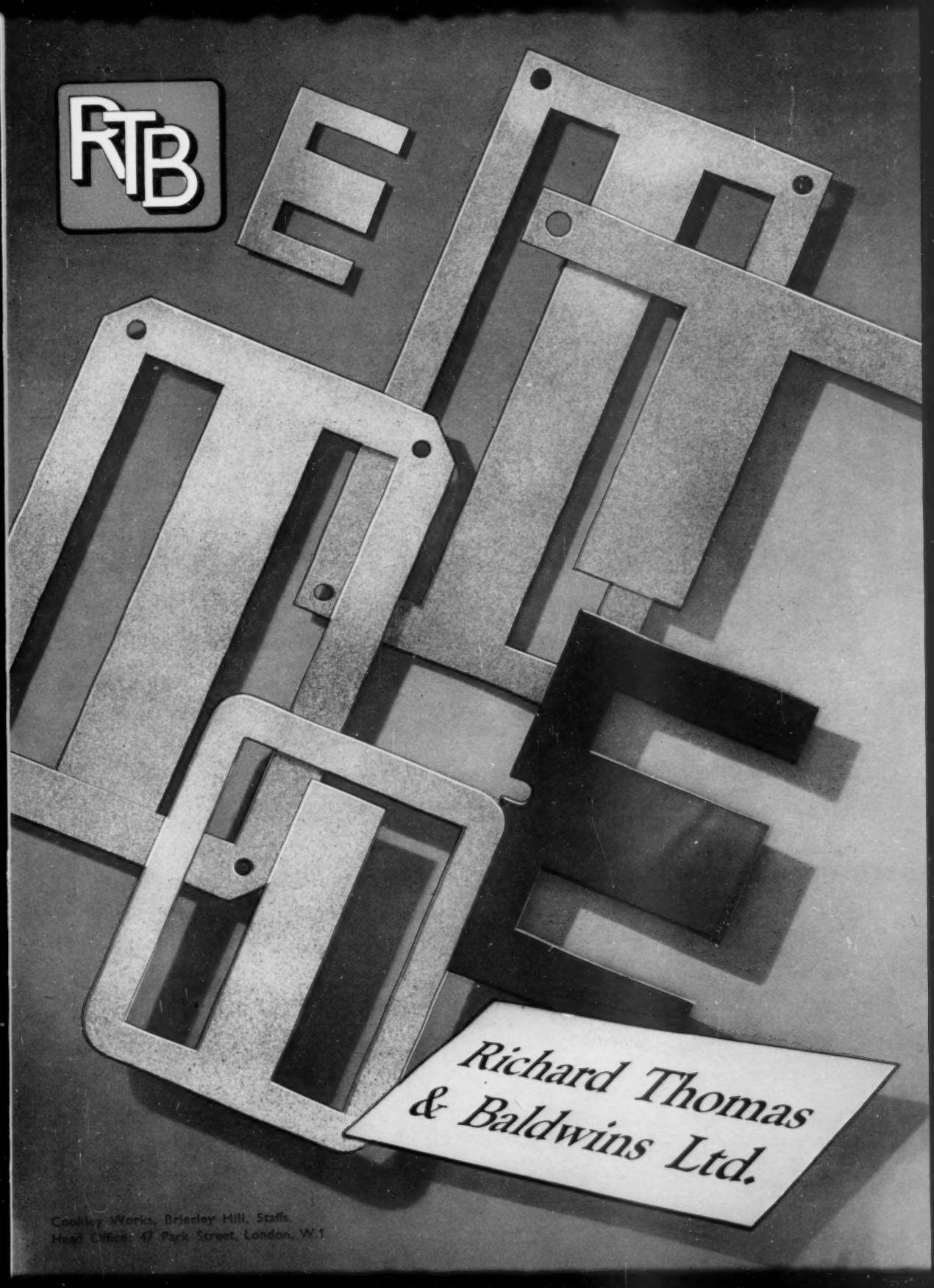
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Cookley Works, Brierley Hill, Staffs.
Head Office: 47 Park Street, London, W.1

Polypole CABLE COUPLER

For- *specialised remote control—centimetre radio links—ground radar—outside broadcast television*

BICC Couplers and Cables are intended for the outdoor inter-connection of equipment, such as that mentioned above. Each application calls for composite trailing cables containing both R.F. units and other polythene insulated conductors.

BICC Polypole Mark III Couplers are available in two versions, designed for use with two standard types of BICC outdoor trailing cables. The Mark IIIA cable and coupler incorporates three coaxial circuits, and the Mark IIIB three screened twin circuits. In addition, both cables contain three triplets and 21 other conductors.

The couplers are permanently moulded to the ends of the cable in the factory. This technique provides a remarkably robust coupler which is virtually free from the hazards of conductor breakages near to, or within the coupler.

If you are interested in the uses of BICC Polypole Cable Couplers, we will be pleased to send you further information.

Note these important features

1. The couplers are assembled with the conductors in tension to ensure that they each contribute their share of the total strength.
2. Polythene injection moulding permits a watertight assembly.
3. Screwed lock rings provide forced engagement and withdrawal. The overall metal housing can also be easily replaced should it become damaged.
4. The cable itself is designed with a symmetrical cross section to provide the greatest reliability under severe handling.

BI Callender's

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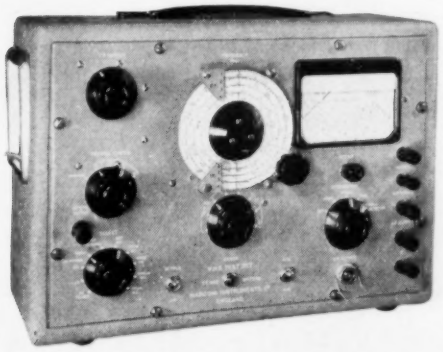
Marconi V.H.F. Test Set Type TF 982

For Mobile Transceivers

For the overall checking of a.m. transmitters and receivers, particularly of the mobile type, the Test Set is primarily a v.h.f./i.f. signal generator with 1,000 c/s modulation facilities; it may also be used as an a.f. and r.f. power meter, a variable-level a.f. source, an r.f. local field detector and a five-range a.c./d.c. test meter.

The signal generator section provides four v.h.f. ranges covering from 60 to 200 Mc/s and four i.f. ranges, each having a bandwidth of 250 kc/s, centred at 1.6, 2.1, 4.86 and 5.0 Mc/s respectively. Facilities are provided for crystal standardisation at the centre frequency of each i.f. range. The v.h.f./i.f. output is derived via a step attenuator and terminating unit giving calibrated outputs from 2μV to 2 mV at 52 and 75 ohms; uncalibrated outputs up to a maximum of 400 mV are also available.



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Signal Generators · Bridges · Valve Voltmeters · Q. Meters · Wavemeters
Frequency Standards · Wave Analysers · Beat Frequency Oscillators

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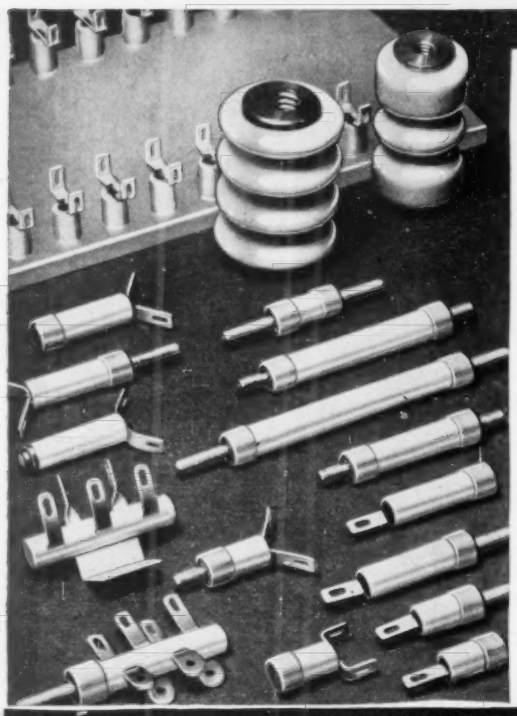

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U.I.C. "Cactus" and "Porcupine" Tags and the almost limitless variety of Stand-off Insulators will solve all your anchorage problems quickly, easily and efficiently.
These latest aids to neat and economical wiring and component assembly are tough, fireproof, and constructed entirely without solder. Only low-loss ceramics and electro-plated brass are used in the manufacture of these best of all anchorages.

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Insulation perfection

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Three-Range **PROBE MILLIVOLTMETER**

Model VM6451

SELF-CONTAINED STABILISED POWER PACK

ACCURACY TO ± 5% OF FULL SCALE DEFLECTION

AUTOMATIC OVERLOAD PROTECTION

Designed to provide direct readings for A.C. voltages from 10 millivolts to 2 volts. Three ranges are provided and readings are accurate within 1db from 50 cycles to 150 megacycles. The double triode measuring valve is detachable from the cabinet. This instrument can also be used for modulation carrier detection with the use of headphones. Write for full specification.

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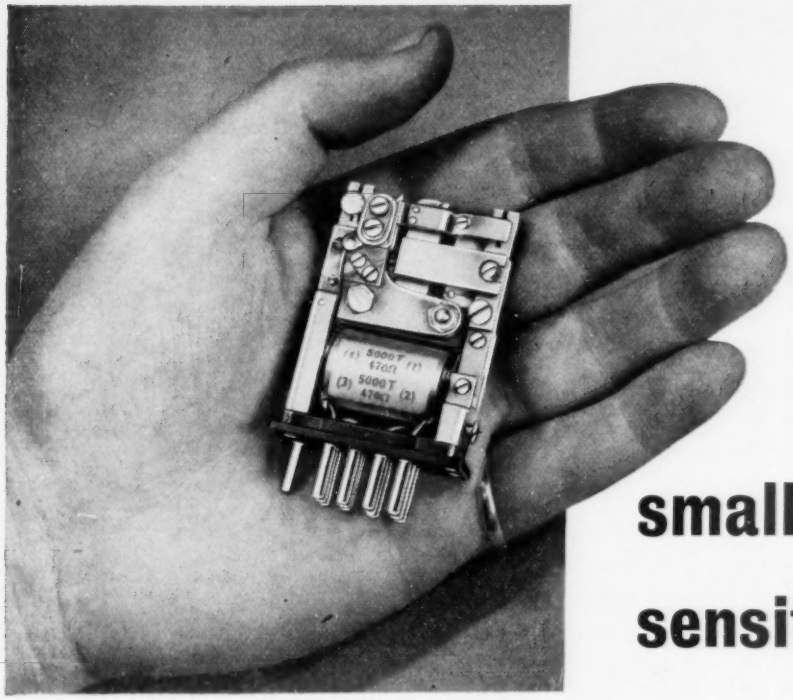
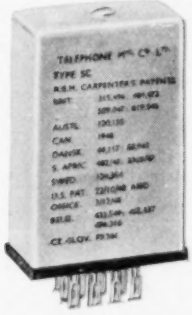


small *and* sensitive

Considering it is only the size of a matchbox the TYPE 5 CARPENTER POLARIZED RELAY is capable of a surprisingly high performance providing answers to problems in many fields of electrical engineering. Outstanding features of the Type 5 include:



HIGH OPERATIONAL SPEED
FREEDOM FROM CONTACT REBOUND
IMMUNITY FROM POSITIONAL ERROR
GOOD CONTACT PRESSURES
HIGH SENSITIVITY
ACCURACY OF SIGNAL REPETITION
RUGGED DESIGN
EXCEPTIONAL THERMAL STABILITY

Plug or solder tag base optional.

Dimensions — (*With cover, excluding connecting pins.*) 2 $\frac{3}{16}$ *ins. high.* 1 $\frac{7}{16}$ *ins. wide.* $\frac{3}{4}$ *in. deep.* Weight (*including socket*) 4.8 *oz.* (137 *gm.*). *Complete specification and further details of the complete range of Carpenter Relays may be had on request.*

TMC CARPENTER POLARIZED RELAY TYPE 5

Manufactured by the Sole Licensees:

TELEPHONE MANUFACTURING CO. LTD

Contractors to Governments of the British Commonwealth and other Nations

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G.E.C.

QUARTZ CRYSTAL UNITS

for Long Term Stability

Q.C. 327

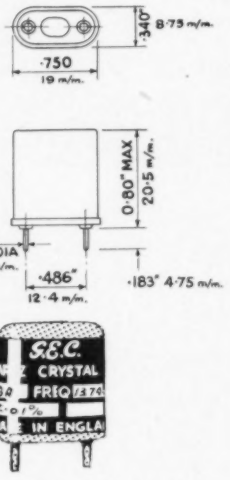
Frequency range 6,000 Kc/s to 16,000 Kc/s

Tolerance ± 0.01% or ± 0.005%

Type BA,
frequency change not exceeding 0.01% from 0°C to +70°C

Type DA,
frequency change not exceeding 0.01% from −30°C to +45°C

Type EA,
frequency change not exceeding 0.002% from +65°C to +80°C



Gross weight 3.5 ozs (99 gms)
Nett weight 0.15 oz (4.25 gms)

This style holder corresponds to Inter Services style D and the U.S. style H.C./6 U.

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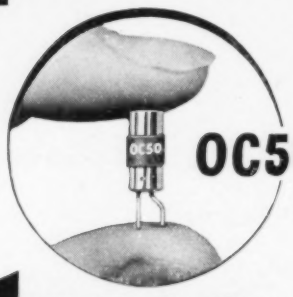
transistors for circuit experiments



OC50

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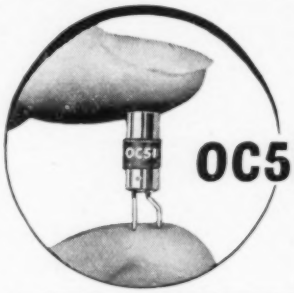
● *The OC50 and OC51 are readily available for experimental purposes at a price comparable with that of subminiature valves.*



OC51

Point-contact transistor type No.		OC50	OC51
Max. negative collector-to-base voltage	(V)	30	50
Max. collector current	(mA)	−12	−15
Max. dissipation	(mW)	120	100
Max. emitter current	(mA)	10	12
Max. ambient temperature	(°C)	40	55
Max. frequency (for α 3db down)	(Mc/s)	1.0	1.5
"Full-on" voltage across transistor		★	
Collector "cut-off" current			★

★ Superior type for this characteristic

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WIRELESS ENGINEER

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Constant-H.T.-Current Amplifier

WHEN an amplifier is required to have a response extending down to zero frequency, difficulty often arises because of the internal impedance of the h.t. supply. In some cases this impedance must be made negligibly small through the use of a suitably stabilized supply, but a well-known alternative is to design each stage of the amplifier so that it draws only a constant current from the h.t. supply. The usual way of doing this is by using push-pull stages.

It is not always realized that there is an alternative to push-pull. Two valves, whose changes of current are equal and opposite are still needed, but they are in cascade, not in push-pull. The circuit is shown in Fig. 1 in its simplest form and comprises merely a triode amplifier directly-coupled to a cathode follower. An increase of current i_{a1} in V_1 increases the voltage drop across R_1 and so changes the grid potential v_2 of V_2 negatively. This reduces the current i_{a2} in V_2 and the output voltage v_0 across R_2 changes negatively. If the fall of current in V_2 equals the rise of current in V_1, the total current drawn by the two valves is unaltered, and any impedance in the h.t. supply has no effect.

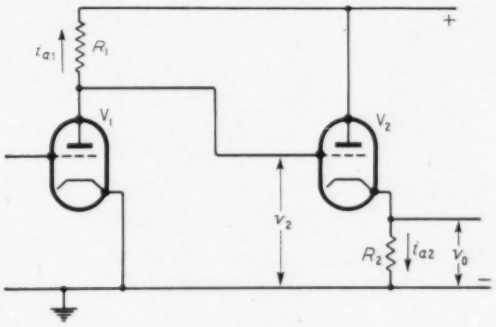


Fig. 1.

If the cathode follower were perfect so that the change of output voltage equalled the change of grid voltage, the only requirement for the constancy of total anode current would be the very simple one of making R_1 and R_2 equal. Equal changes of voltage across equal resistances require equal changes of current in them. In practice, the cathode follower is not perfect and, as is well known, it has the 'amplification'

$$A_2 = \frac{A}{1 + A} \quad \text{where } A = \frac{g_{m2} r_{a2} R_2}{r_{a2} + R_2},$$

the basic amplification without the cathode feedback.

The anode current of V_2 is $i_{a2} = v_0/R_2$ and the grid voltage is $v_2 = v_0/A_2 = i_{a2}R_2/A_2$. The anode current of V_1 is $i_{a1} = v_2/R_1$, hence

$$i_{a1} = i_{a2} \frac{R_2}{A_2 R_1}$$

The anode currents are thus equal, and the supply current is zero, when

$$R_1 = R_2 (1 + 1/A)$$

In the circuit of Fig. 1, the attainment of the proper operating conditions for V_2 demands that the anode voltage of V_1 be less than the cathode voltage of V_2. These proper conditions are easily found if the two valves are alike and if the mutual conductance is high enough for the amplification A_2 of V_2 to be negligibly different from unity.

Then R_1 and R_2 are equal, both valves require the same grid–cathode bias and they each pass the same standing anode current. For the grid of V_2 to be negative to its cathode, the voltage drops across R_1 and R_2 (or, since the drops across R_1 and R_2 are equal, twice the voltage drop across R_1) must equal the h.t. supply voltage plus the required bias.

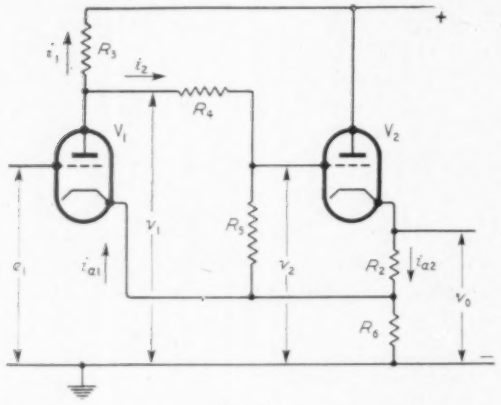


Fig. 2.

To a first approximation the grid bias is independent of the load and is about one-third of the way between the limiting values set by grid current on the one hand and anode-current cut-off on the other. From the valve curves, and the known h.t. supply voltage, therefore, we can choose a grid-voltage curve upon which the operating point must lie; twice the voltage drop across R_1 must equal the supply voltage plus the bias voltage, so we can calculate the voltage drop across R_1. The anode–cathode voltage is the h.t. supply voltage less this voltage drop and so we find the operating point. The load line can be drawn in and R_1 and R_2 evaluated.

The resulting values for R_1 and R_2 are sometimes too high for other requirements, such as high-frequency response, and the modified circuit of Fig. 2 permits more latitude in the choice of operating conditions. In this circuit, the anode current of V_1 divides between R_3 and $R_4 + R_5$ and it is only the part in R_3 which flows in the h.t. supply. If $n = R_5/(R_4 + R_5)$, we have

$$v_2 = nv_1 = ni_1R_3 = i_{a2}R_2/A_2$$

and for equality of i_1 and i_{a2}

$$R_3 = \frac{R_2}{n}\left(1 + \frac{1}{A}\right)$$

So far, R_6 has not been considered. This is a bias resistor for V_1. With the connection of R_5 to the cathode of V_1, only the supply current flows in R_6 and its presence there has no effect on the signal conditions.

It appears at first as if the presence of the potential divider R_4, R_5 would reduce the amplification. This is not necessarily the case, however. If the voltages and currents of V_2 are kept constant, i_1 must be constant whatever the value of n. But $v_2/i_1 = R_3n$ and is a constant. Now the amplification of V_1 is

$$A_1 = \frac{v_1}{e_1} = \frac{\mu_1 n \dfrac{R_3(R_4 + R_5)}{R_3 + R_4 + R_5}}{r_{a1} + \dfrac{R_3(R_4 + R_5)}{R_3 + R_4 + R_5}}$$

substituting for R_3 and re-arranging, we get

$$A_1 = \frac{g_{m1}v_2/i_1}{1 + \dfrac{v_2/i_1}{n}\left(\dfrac{1}{r_{a1}} + \dfrac{1}{R_4 + R_5}\right)}$$

If $r_{a1} \to \infty$, as with a pentode, $A_1 \to g_{m1}\, v_2/i_1$ and is independent of n. If $r_{a1} \to 0$, $A_1 \to \mu_1 n$. The normal triode condition is intermediate between these extremes, and the use of the potential divider does result in some decrease of amplification.

As shown, the circuit is best suited to low-frequency operation where triodes are suitable. If the stage is also to work at higher frequencies, complications ensue because of valve and other stray capacitances. A pentode becomes necessary for V_1 to avoid Miller effect and a compensated coupling is needed between the valves so that the circuit takes the form shown in Fig. 3.

For the purposes of calculation we can let $R_{3a} + R_{3b} = R_3$ of Fig. 2 and ignore C_1 and C_2, if we make $R_{3a}/R_{3b} = R_5/R_4$ and $C_1R_4 = C_2R_{3b}$, the usual relation for this form of compensated

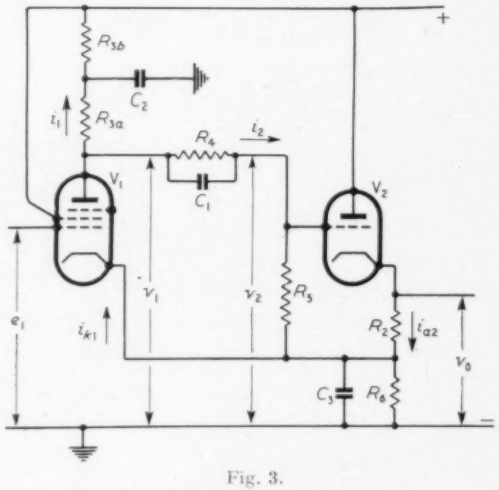


Fig. 3.

coupling. The load on V_1 falls with increasing frequency but the effective potential-divider ratio changes equally and for a given current from V_1 the grid voltage of V_2 is constant at all frequencies.

A discrepancy arises because of the screen current of V_1. To correct for this R_3 must be increased in the ratio i_{k1}/i_{a1}. If a screen dropping resistor is used for V_1 the matter becomes still more complicated. It must not be by-passed and feedback occurs because of it. Because of valve capacitance the feedback will change and Miller effect may occur at high frequencies. It is, therefore, desirable to feed the screen from a constant-voltage point as shown in Fig. 3.

While the proper conditions for balance are easily determined, as has been shown, it is a nice problem in design to choose values for the resistors which adhere to the balance relations and, at the same time, give the proper operating conditions for the valves.

It is to be noted as a great merit of the circuit that, because of the heavy negative feedback in V_2, this stage is very nearly linear, hence the anode-current waveforms of the two valves must be very nearly alike. Distortion in V_1 does not affect the matter and the characteristics of V_1 do not affect the balance. Because of the feedback, large changes in the characteristics of V_2 have only a small effect on the balance and, in practice, it is resistor values only which have a large effect. For good and well-maintained balance, close-tolerance and stable resistors must be used.

When V_1 is a pentode, a change of valve may affect the balance somewhat, since the ratio of anode to screen currents may not be exactly the same for all specimens. W. T. C.

TRANSMISSION-LINE MATCHING SYSTEM

Properties of Two Quarter-Wave Transformers

By R. E. Collin, B.Sc., and J. Brown, M.A., A.M.I.E.E.

(Department of Electrical Engineering, Imperial College of Science and Technology, London)

SUMMARY.—The two-section quarter-wave transformer is analysed in detail. It is shown that by tolerating a certain mismatch at the centre of the band it is possible to obtain a perfect match at two frequencies located symmetrically about the centre frequency. This arrangement gives an increase of 20–45% in bandwidth over the commonly-used binomial transformer. The binomial transformer is shown to be a special case of the above transformer, obtained when the two frequencies at which a perfect match occurs coincide at the centre frequency. A simple design procedure and performance curves are also given.

Introduction

WHEN a transmission line is to be terminated in a resistive load differing from the characteristic impedance of the line, it is usual to use an intermediate section of transmission line, a quarter of a wavelength long and of characteristic impedance equal to the geometric mean of the load resistance and the characteristic impedance of the main line. For such a single-section quarter-wave transformer a perfect match is obtained at one frequency only and the overall bandwidth is usually small. If a wider bandwidth is required a two-section matching transformer is used. It is usual to arrange the impedances in a binomial progression[1,2], and the two-section transformer is then termed a binomial transformer. Thus $Z_1 = R^{\frac{1}{4}}$, $Z_2 = R^{\frac{1}{4}}$, where Z_2 is the characteristic impedance of the section of transmission line adjacent to the load, R is the load impedance and all impedances are normalized with respect to the characteristic impedance of the main line. It is usually thought that this arrangement provides the maximum bandwidth for the match which can be obtained from two transformer sections. No proof of this has, however, been given and in seeking such a proof it has been found that an alternative arrangement can give a greater bandwidth.

The binomial transformer corresponds to a maximally-flat band-pass filter in filter theory. By tolerating a certain mismatch at the centre frequency it is possible to match a two-section quarter-wave transformer at two frequencies located symmetrically on either side of the centre frequency. This corresponds to the case of an overcoupled circuit and gives a much wider usable bandwidth. It is found that from 20% to 45% increase in bandwidth over the binomial transformer can usually be obtained. In this paper the optimum bandwidth two-section transformer is analysed in detail. The design formulae are given and a comparison with the two-section binomial transformer is made. This technique has

MS accepted by the Editor, June 1953.

been applied to the 4-window directional coupler used in waveguide work.[3]

The circuit to be studied is shown in Fig. 1, in which the resistive load R is to be matched to a transmission line whose characteristic impedance is taken for convenience as unity. The matching section consists of the two sections of transmission line 1 and 2, whose lengths are L_1 and L_2 respectively, with characteristic impedances Z_1 and Z_2. The problem being considered is the determination of the four parameters L_1, L_2, Z_1 and Z_2 so that the complete termination is as well matched as possible over the widest possible frequency range. More precisely, the parameters will be calculated to maximize the bandwidth within which the standing-wave ratio does not exceed a specified value, S_0. It is found to be extremely difficult to tackle the problem in its complete generality and one restriction will be imposed; i.e.,

$$\frac{L_1}{V_1} = \frac{L_2}{V_2} = p, \quad \ldots \quad (1)$$

where V_1, V_2 are respectively the phase velocities for lines 1 and 2. The phase velocities and the various impedances will be assumed independent of frequency over the frequency band for which the system is to be matched. In view of Equ. (1) the phase-shift across either line 1 or line 2 will be

$$\theta = 2\pi fp \quad \ldots \quad (2)$$

for any frequency f. The phase angle θ is therefore directly proportional to frequency.

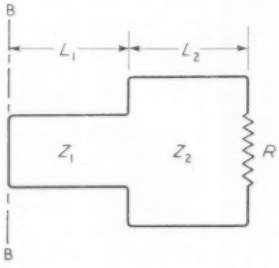


Fig. 1. *Two sections of line of length L_1 and L_2 and normalized impedances Z_1 and Z_2 used to match a normalized load R.*

The input impedance of the structure to the right of B in Fig. 1 can be calculated from ordinary transmission-line theory and is found to be

$$Z_{in} = \frac{Z_1R(Z_2 - Z_1t^2) + jZ_1Z_2(Z_1 + Z_2)t}{Z_2(Z_1 - Z_2t^2) + jR(Z_1 + Z_2)t} \quad \ldots (3)$$

where $t = \tan\theta = \tan 2\pi fp$.

The complex reflection coefficient ρ at the plane B is given by

$$\rho = \frac{Z_{in} - 1}{Z_{in} + 1} \quad \ldots \quad (4)$$

The modulus of the reflection coefficient r, can now be calculated as a function of frequency by substituting in the above equation for Z_{in}. The problem reduces to determining the values of Z_1, Z_2 and p so that the standing-wave ratio, S, is less than S_0 over as wide a range of frequencies as possible. When stated in this form the problem becomes very similar to that of designing a flat-topped band-pass filter and so a similar procedure can be used. Inspection of the expression for Z_{in} shows that it is a symmetrical function of frequency about the frequency $1/4p$ at which $\tan\theta$ becomes infinite. This suggests that $1/4p$ be made equal to f_0, the mid-frequency of the band for which the system is to be matched. Then

$$\theta = 2\pi fp = \pi f/2f_0 \quad \ldots \quad (5)$$

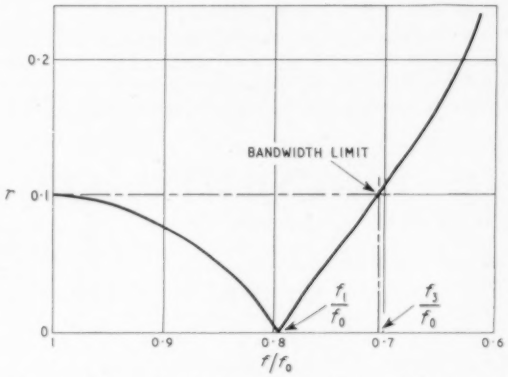


Fig. 2. *Form of reflection-coefficient curve; this curve is drawn for $R = 5$, $r_0 = 0{\cdot}1$.*

It is possible to select Z_1 and Z_2 so that ρ becomes zero for the midband frequency f_0, the necessary condition being $Z_1^2R = Z_2^2$. For a maximum bandwidth this gives the binomial transformer, as will be shown later. A wider band can be achieved by making r zero at some frequency f_1 less than f_0. From the symmetry of the expression for r with respect to frequency it follows that r will also be zero at a frequency f_2 equal to $2f_0 - f_1$.

Let θ_1 be the value of θ corresponding to f_1 so that $\theta_1 = \pi f_1/2f_0$. The condition that r equals zero is that Z_{in} must be unity so that from Equ. (3)

$$Z_1R(Z_2 - Z_1t_1^2) = Z_2(Z_1 - Z_2t_1^2)$$

and

$$Z_1Z_2(Z_1 + Z_2)t = R(Z_1 + Z_2)t$$

where $t_1 = \tan\theta_1$.

The above equations give

$$Z_1Z_2 = R \quad \ldots \quad (6)$$

$$t_1^2 = \frac{Z_1Z_2(R-1)}{Z_1^2R - Z_2^2} = \frac{R(R-1)}{Z_1^2R - R^2/Z_1^2} = \frac{Z_1^2(R-1)}{Z_1^4 - R} \quad \ldots \quad (7)$$

This provides a unique relation between Z_1 and

Z_2 so that if Z_1 is found, Z_2 follows immediately. The next step is to select a suitable value of the frequency f_1, for which the reflection coefficient is made zero. An examination of the frequency dependence of the reflection coefficient shows that it is always of the form shown in Fig. 2, there being a maximum value of r at the midband frequency f_0. As the frequency approaches f_0, t tends to infinity and so from Equ. (3) Z_{in} becomes the pure resistance Z_1^2R/Z_2^2; i.e., Z_1^4/R at the frequency f_0. The reflection coefficient is therefore

$$r' = \frac{Z_1^4 - R}{Z_1^4 + R} \quad .. \quad .. \quad .. \quad .. \quad (8)$$

It is convenient to restrict the analysis at this point to the case when R exceeds unity, so that $Z_1^4 - R$ is positive, as may be seen from Equ. (7) since t_1^2 must be positive. The modification required when R is less than unity can easily be made to the remainder of the analysis. When R exceeds unity, r' in Equ. (8) is a positive real quantity and equals the modulus of the reflection coefficient, r'. It appears reasonable from Fig. 2 to suppose that the greatest bandwidth will be obtained if the reflection coefficient at f_0 is made equal to the maximum tolerable reflection coefficient r_0. This result can be justified analytically. The required value of Z_1 is now obtained by letting r' equal r_0 when from Equ. (8)

$$Z_1^4 = R\,\frac{1 + r_0}{1 - r_0} = RS_0 \quad .. \quad .. \quad (9)$$

where S_0 is the corresponding voltage standing-wave ratio (taken greater than unity).

Also from (6)

$$Z_2 = \frac{R}{Z_1} = \frac{R^{3/4}}{S_0^{1/4}} \quad . \quad .. \quad .. \quad .. \quad (10)$$

Substitution for Z_1 and Z_2 in Equ. (3) gives

$$\begin{aligned} Z_{in} &= \frac{R^{5/4}S_0^{1/4}(R^{3/4}/S_0^{1/4} - R^{1/4}S_0^{1/4}t^2) + jR(R^{1/4}S_0^{1/4} + R^{3/4}/S_0^{1/4})t}{R^{3/4}/S_0^{1/4}(R^{1/4}S_0^{1/4} - R^{3/4}/S_0^{1/4}t^2) + jR(R^{1/4}S_0^{1/4} + R^{3/4}/S_0^{1/4})t} \\ &= \frac{R^{3/2}S_0^{1/2}(R^{1/2} - S_0^{1/2}t^2) + jR^{5/4}S_0^{1/4}(S_0^{1/2} + R^{1/2})t}{R(S_0^{1/2} - R^{1/2}t^2) + jR^{5/4}S_0^{1/4}(S_0^{1/2} + R^{1/2})t}. \end{aligned}$$

Hence
$$\rho = \frac{R(R-1)\sqrt{S_0} - R^{3/2}(S_0 - 1)t^2}{R(R+1)\sqrt{S_0} - R^{3/2}(S_0 + 1)t^2 + 2jtR^{5/4}S_0^{1/4}(\sqrt{S_0} + \sqrt{R})},$$

and
$$r^2 = |\rho|^2 = \frac{[(R-1)\sqrt{S_0} - \sqrt{R}(S_0 - 1)t^2]^2}{[(R+1)\sqrt{S_0} - \sqrt{R}(S_0 + 1)t^2]^2 + 4\sqrt{S_0}(\sqrt{S_0} + \sqrt{R})^2\sqrt{R}t^2} \quad . \quad .. \quad .. \quad (11)$$

The bandwidth of the matched system is obtained from the frequency f_3 for which r again equals r_0 (see Fig. 2). If the corresponding value of t is t_3, then from (11)

$$\begin{aligned} &(S_0 - 1)^2[(R+1)\sqrt{S_0} - \sqrt{R}(S_0 + 1)t_3^2]^2 \\ &\quad + 4\sqrt{S_0}\sqrt{R}(S_0 - 1)^2(\sqrt{S_0} + \sqrt{R})^2t_3^2 \\ &= (S_0 + 1)^2[(R-1)\sqrt{S_0} - \sqrt{R}(S_0 - 1)t_3^2]^2 \end{aligned}$$

since

$$r_0 = \frac{S_0 - 1}{S_0 + 1}.$$

This equation simplifies to a linear equation in t_3^2 whose solution is

$$2t_3^2 = \frac{R-1}{\sqrt{R}}\,\frac{\sqrt{S_0}}{S_0 - 1} - 1, \quad .. \quad .. \quad (12)$$

The frequency f_3 which defines the lower limit of the frequency band within which $r \leqslant r_0$ is given by $2f_0/\pi \tan^{-1}t_3$. There is a very simple relation between t_3 and t_1, the value of $\tan\theta$ corresponding to the frequency f_1 at which r becomes zero. From Equs. (7) and (9)

$$t_1^2 = \frac{R-1}{\sqrt{R}}\,\frac{\sqrt{S_0}}{(S_0 - 1)}, \quad .. \quad .. \quad (13)$$

and hence from Equ. (12) it follows that

$$t_3^2 = \tfrac{1}{2}(t_1^2 - 1) \quad . \quad .. \quad .. \quad .. \quad (14)$$

The Binomial Transformer

A two-section binomial transformer may be designed in one of two ways.[2] Only one need be considered and its properties are most easily obtained as a special case of the previous analysis. Suppose the two frequencies at which the reflection is zero are made to become closer together, eventually coinciding at the design frequency f_0. It follows at once from equation (7) that

$$Z_1 = R^{1/4},\ Z_2 = R^{3/4}.$$

The reflection coefficient has a double zero at the design frequency and so its first derivative with respect to frequency will also be zero. The binomial transformer so obtained has therefore a similar behaviour to a maximally-flat band-pass filter. The bandwidth of this arrangement is obtained, as before, by finding the frequency for which the reflection coefficient equals r_0.

For the binomial transformer

$$Z_1 = R^{1/4},\ Z_2 = R^{3/4} \text{ and therefore}$$

$$Z_{in} = \frac{R\,(R - \sqrt{R}t^2) + jR\,(R^{1/4} + R^{3/4})t}{R - R^{3/2}t^2 + jR\,(R^{1/4} + R^{3/4})\,t}.$$

$$\text{with } \rho = \frac{Z_{in}-1}{Z_{in}+1} = \frac{R(R-1)}{R^2+R-2R^{3/2}t^2+2jR^{5/4}(1+\sqrt{R})t}.$$

The corresponding value of r^2 is given by

$$r^2 = |\rho|^2 = \frac{(R-1)^2}{(1-2\sqrt{R}t^2+R)^2+4\sqrt{R}(1+\sqrt{R})^2t^2}.$$

$$= \frac{(R-1)^2}{(1+R)^2-4\sqrt{R}t^2(1+R-1-2\sqrt{R}-R)+4Rt^2}.$$

When $r = r_0$ let $t = t_B$, where t_B defines the bandwidth limit of the binomial transformer, and it follows that

$$4R(t_B^2+1)+(R-1)^2 = \frac{(R-1)^2}{r_0^2}$$

and therefore

$$t_B^2 = t_1^2 - 1 \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (15)$$

From the above equation it is seen that there is a unique relation between the bandwidth of the binomial transformer and that based on the alternative design given above. The improvement resulting from the use of this design is illustrated

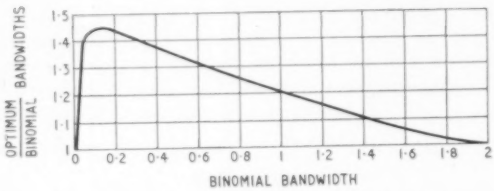


Fig. 3 (above). *This curve shows the ratio of the bandwidths obtainable with the two forms of matching transformer.*

Fig. 4 (right). *Optimum fractional bandwidth as a function of R for various standing-wave ratios S_0.*

in Fig. 3, in which the ratio of optimum bandwidth to that of the binomial bandwidth is plotted against binomial bandwidth. It is seen that a maximum improvement of about 45% is possible. In practice an increase of 30% or more is usually obtained. In this paper the bandwidth is expressed as a fraction of the centre frequency f_0, so that for the above designed transformer the bandwidth is taken as $2\frac{f_0-f_3}{f_0}$.

Design Procedure

Case 1: Maximum Value of Tolerable Voltage Standing-Wave Ratio given

When S_0, the maximum tolerable voltage standing-wave ratio is given then Z_1 is determined by Equ. (9) as $Z_1 = (RS_0)^{1/4}$.

The corresponding value of Z_2 is found from the relation $Z_2 = R/Z_1 = R^{\frac{3}{4}}/S_0^{\frac{1}{4}}$.

The bandwidth is determined from Equs. (13) and (14) which, when combined, give

$$t_3^2 = \tfrac{1}{2}\left[\frac{R-1}{\sqrt{R}}\frac{\sqrt{S_0}}{S_0-1}-1\right]$$

$$\text{Fractional bandwidth} = 2\frac{f_0-f_3}{f_0}$$

where $f_3 = \frac{2f_0}{\pi}\tan^{-1}t_3$.

Case 2: Required Bandwidth given

If the required bandwidth is given then f_3 is fixed and this in turn fixes t_3 and t_1.

Thus $t_3 = \tan\frac{\pi f_3}{2f_0}$ and from Equ. (14)

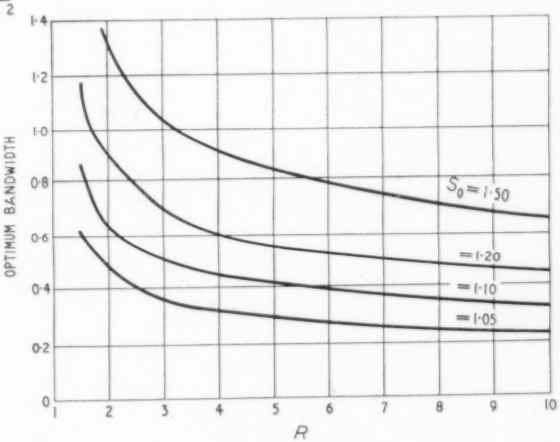
$t_1^2 = 2t_3^2 + 1$.

This value of t_1^2 can be used in equation (13) to obtain the value of S_0, thus

$$\frac{S_0}{(S_0-1)^2} = \frac{R}{(R-1)^2}t_1^4 = a,$$

and hence

$$S_0 = \left(1+\frac{1}{2a}\right)+\sqrt{\left(\frac{2a+1}{2a}\right)^2-1}$$



Since S_0 must exceed unity, only the positive sign is applicable.

Alternatively Z_1 may be obtained from Equ. (7) as

$$Z_1 = \left[\sqrt{\left(\frac{R-1}{2t_1^2}\right)^2+R} - \frac{R-1}{2t_1^2}\right]^{1/2}.$$

The corresponding value of Z_2 is R/Z_1.

The above analysis is for the case when R exceeds unity. When R is less than unity the only alteration required in the above design formulae is to replace R by $1/R$, Z_1 by $1/Z_1$ and Z_2 by $1/Z_2$.

In Fig. 4 is plotted a family of curves giving the optimum fractional bandwidth as a function of R for four values of S_0. The performance for any particular case can be obtained from these curves.

Conclusion

It has been shown that the binomial transformer does not give the maximum usable bandwidth. A simple design procedure has been worked out whereby a two-section transformer can readily be designed if the maximum tolerable voltage standing-wave ratio or required bandwidth is given. Although the lengths of the transformer sections were chosen as a quarter-wave long at the frequency f_0, it does not seem logical to expect any significant improvement in bandwidth by staggering the lengths since it is only possible to make ρ zero at two frequencies. There are four adjustable parameters and each zero requires two, one to adjust the modulus of Z_{in} to unity and one to bring the phase angle of Z_{in} to zero.

The design procedure which has been illustrated in this paper may obviously be extended to a matching network using n quarter-wave transformers. The reflection coefficient can then be made zero at n different frequencies.

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EXPERIMENTAL INVESTIGATION OF GRID NOISE

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SUMMARY.—Results are given of a detailed investigation of triode noise factor, with particular reference to correlation of induced grid noise with shot noise. It is deduced that correlation is very slight and that, although the optimum value of noise factor can be calculated fairly accurately from the values of shot noise and optimum source resistance, the latter must be found by experiment and therefore the theory is not of major practical importance.

1. Introduction

THE first published papers[1,2] on this subject appeared almost simultaneously, and since then there have been few papers published which give satisfactory experimental evidence of the validity of the theories. The possibilities of phase-correlation effects were first brought to the notice of one of us (N. Houlding) in 1944 by an unpublished paper by Peterson, who later published a different paper[3] on his theories of methods of compensating noise fluctuations by transit-time effects. Preliminary experiments in 1944 on detuning the input stage of a common-grid circuit were disappointing, but experiments with common-cathode triode circuits showed that the first result was due to a fortuitous optimization of the input stage with the practical method of adjustment adopted.[4] A series of experiments was made in 1946-7 with the object of gaining useful practical data, and possibly throwing further light on the subject.

Since that date, other papers on transit-time noise effects and the correlation with shot noise have appeared.[5,6,7] An excellent paper with a clear physical explanation is that by W. A. Harris,[8] which will probably become as widely referred to as the series by Thompson, North and Harris,[9] although his tables of comparative performance with different valves are not backed by experiment.

The outstanding difficulty in an exact analysis is the calculation of the effect for the whole range of different velocity classes of electrons. A critical discussion of the problem is given in a paper by Campbell, Francis and James,[10] who point out that the existing formula for induced grid noise is dependent on valve theory which does not take transit-time effects into account. In fact, the phenomena of space-charge limitation and space-charge smoothing only occur because of the different velocities of emission, and it is remarkable that so much progress has been made with theories which neglect the velocity distribution.

The results of our experiments were in disagreement with the simplified theory, in particular in respect of correlation effects, and since papers by

MS accepted by the Editor, March 1953

Bell[7] and A. I. Harris[11] have appeared, we decided to publish our evidence.

In what follows, it is perhaps necessary to apologize to some readers for the treatment in which all noise sources are referred to the grid, sometimes with 'transformation' from other terminals which is based on physical argument rather than rigorous algebra. This is a convention which has become common in analysis of noise, and will be familiar to most, and easier to follow than a rigorous network analysis with cumbersome expressions.

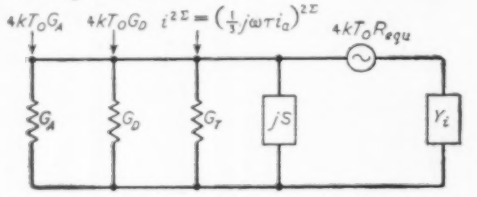


Fig. 1. Equivalent noise circuit of triode; Y_i = input admittance due to feedback. Arrows indicate current generators.

2. Noise Factor with Fully Correlated Induced Grid Noise

From the theory[1] that induced grid noise is given by $\delta i_g = \frac{1}{3}\, j\omega\tau\delta i_a$, where τ is the cathode–grid transit time and δi_g and δi_a are grid and anode current fluctuations, the following expression for noise factor of a triode can be derived. (Fig. 1.)

$$n = 1 + \frac{G_D}{G_A} + \frac{(G_A + G_D + G_T)^2}{G_A} R_{equ} + \frac{(S + \frac{1}{3}\,\omega\tau\, g_m)^2}{G_A} R_{equ} \quad \ldots \quad (1)$$

where n = single frequency noise factor at mid-band
S = net susceptance between grid and cathode excluding that due to feedback
G_D = input circuit loss conductance at grid
G_A = source conductance at grid
G_T = transit-time input conductance
g_m = mutual conductance (negative)
R_{equ} = shot noise equivalent resistance (See Ref. 4 for terminology)

Hence, when $S = +\left|\frac{1}{3}\,\omega\tau\, g_m\right|$, the noise factor should be given by:—

$$n = 1 + \frac{G_D}{G_A} + \frac{(G_A + G_D + G_T)^2}{G_A} R_{equ} \quad (2)$$

and, the noise factor will have an optimum value given by:—

$$n_{opt} = 1 + 2R_{equ}(G_D + G_T) + 2G_0R_{equ} \quad \ldots \quad (3)$$

with

$$G_A{}^2 = G_0{}^2 = \frac{G_D}{R_{equ}} + (G_D + G_T)^2 \text{ or } G_0 \approx \sqrt{\frac{G_D}{R_{equ}}}$$

The theoretical detuning required is, on some arguments, equal to the capacitance change due to space-charge (though this is critically discussed by A. I. Harris[11]). Hence the net result would be that the space-charge capacitance effect on noise factor is exactly similar to any other susceptance effects due to feedback. In other words, correct tuning of the input is obtained provided $g_m = 0$. The analogy between transit-time effects on noise and input admittance, and external feedback effects on noise and input admittance, particularly cathode-lead effects, is very appealing in its simplicity.

3. Outline of the Experiments

All our preliminary experiments were made with hope of finding full correlation of induced grid noise because of the extreme importance of such a result, if it could be applied to amplifiers at, say, 1,000 Mc/s. This is important to state, because the subtleties of bias in experimental work are often more than many of us like to confess.

When it was found that detuning the input circuit did not give the full improvement in noise factor predicted by simple theory, the emphasis was changed to try and establish reliable experimental data from which noise factor could be calculated. The theoretical expression[1,2] for induced grid noise in terms of transit time conductance G_T, namely*

$$i_g{}^{2\Sigma} = \frac{80}{9}\left(1 - \frac{\pi}{4}\right) 4kT_cG_T$$

(where T_c = cathode temperature)

is of little practical use, because G_T is seldom known with any accuracy, and the separation of G_T from the effects of cathode-lead inductance is too difficult. Above all, the correlation with shot noise is not disclosed.

A direct measurement of $i_g{}^{2\Sigma}$ is also fraught with experimental error due to the effects of lead inductances, and although it should be possible to make an accurate measurement of the noise temperature of the total input conductance, this would still be referring back the problem to a separation of the component parts of the input conductance even though the noise temperature of the conductance due to the cathode lead could be calculated.

It was decided, therefore, to measure the noise factor of a triode as a function of frequency and input circuit conditions. Unfortunately, no way out of the dilemma caused by valve lead inductances was found, and it is believed that they play a part in the transformation of the external admittances. It was argued that errors due to the transformation in source admittance would be less than the errors working the reverse way (i.e.,

* $i_g{}^{2\Sigma}$ is the mean-square grid current.

measuring i_g^{2}, etc.), since theoretically the grid-cathode source admittance *excluding effects of feedback* (i.e., effects of anode current in the cathode lead), is the admittance determining the noise factor. Taking as typical values, $L = 0{\cdot}01$ μH, $C = 10$ pF, the transformation should give no more than 2% effect on source admittance. The possibility of working with resonant leads was discarded on the grounds of uncertainty. In addition to valve leads, the lead lengths used in wiring and measurement may cause significant error.

A very careful investigation with pentodes was ruled out because of the extreme complexity of the noise problem with partition noise effects and still further complexities due to lead inductances, but even more because there is no application to low-noise amplifiers.

The methods of measurement have been described in detail in a companion paper.[12] The inductance-neutralized triode circuit was used for the main tests with a CV139 common-grid second stage with an unequal Q coupling to a CV138 pentode feeding the main amplifier-detector. The bandwidth, determined by the last stage of the main amplifier was 120 kc/s.

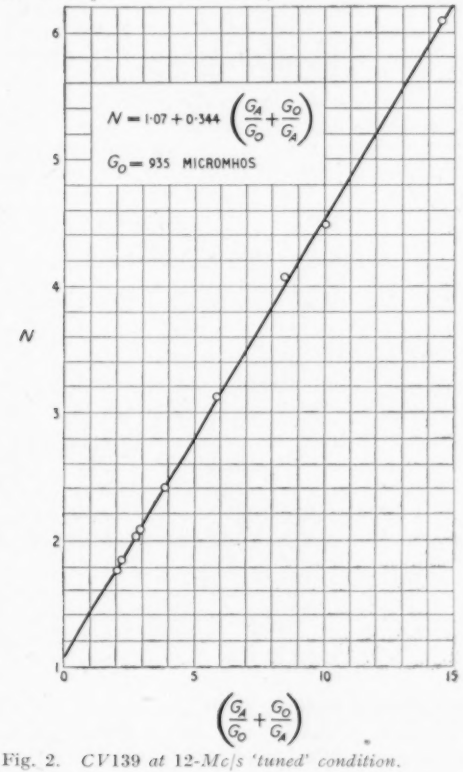


Fig. 2. *CV139 at 12-Mc/s 'tuned' condition.*

The general precautions and corrections outlined in Ref. 12 were made, and in addition a conducting screen was wound round the bulb of the valve and a large negative potential applied to eliminate bulb-charge effects. Input circuit resistances were measured in situ at the grid terminals to offset any change when soldering into the circuit and to minimize errors due to leads used in wiring, and the circuit was operated throughout with the neutralizing coil resonant with the anode–grid capacitance. No particular pains were taken with coil design to achieve high Q coils. The heater voltage was carefully set for all measurements and the results for valve parameters were accurately the same for two separate measurements, within a few weeks, using different observers.

A series of measurements was taken to determine:

(a) Tuned noise factor n_t as a function of source resistance with input circuit resonant including space-charge capacitance.

(b) Detuned noise factor n_d as a function of input circuit susceptance for different values of source conductance.

All the above measurements were made at frequencies of 12, 30, 45 and 60 Mc/s.

The CV139 was chosen for the main investigation because it was known to exhibit larger transit-time effects than the 6AK5, 6J4 or VX3052 (experimental version of the CV408—Osram A1714).

4. Experimental Results

4.1. *Main Programme*

(a) From the measurements in (a) above, it is possible to draw curves of the form

$$n_t = A + BG_A + DG_A^{-1} \quad .. \quad (4)$$

where G_A is the source conductance and A, B and D are constants for any particular valve at one frequency.

If the points are plotted with n_t on linear scale and G_A on logarithmic scale, a symmetrical curve is obtained. The determination of the axis of symmetry, G_0, was improved in accuracy by a check of the equation

$$n_t = A + \sqrt{BD}\left(\frac{G_A}{G_0}+\frac{G_0}{G_A}\right) \quad .. \quad (5)$$

for if G_0 is chosen accurately a plot of n_t against $\left(\frac{G_A}{G_0}+\frac{G_0}{G_A}\right)$ gives a straight line.

(b) From the measurements in (b) above, we can draw curves of the form

$$n_d = E + F(C - C_0)^2 \quad .. \quad (6)$$

where C is the capacitance change from the tuned condition and C_0 is the change for optimum noise

factor. The axis of symmetry was obtained in the simple way, giving more prominence to the points remote from the minimum than those close to. Having determined C_0 a plot of n_d against $(C - C_0)^2$ gives a straight line from which E and F can be found.

These were the graphical methods and Figs. 2, 3 and 4 are examples of the detailed analysis, which was made before calculating the values for R_{equ} from direct measurement to avoid prejudice.

The complete results are given in Tables 1, 2 and 3.

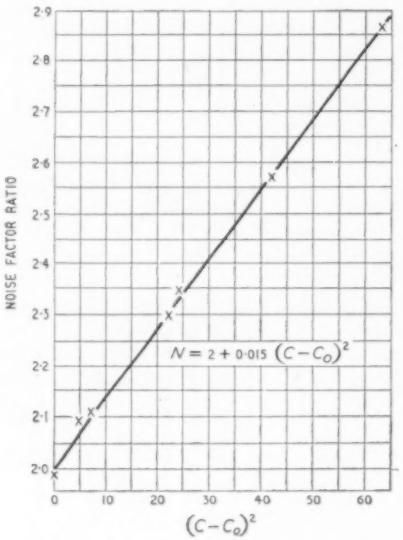
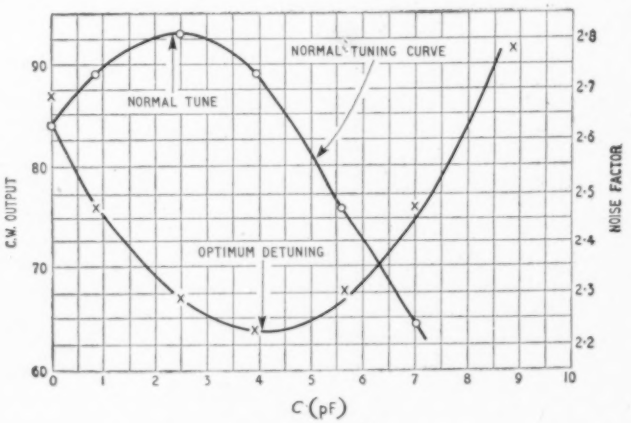


Fig. 3 (above). *CV139 at 45-Mc/s detuning experiment; $G_A = 2095$ micromhos.*

Fig. 4 (right). *CV139 at 60 Mc/s; $G_A = 2390$ micromhos; X = experimental points from which full-line curve $N = 2{\cdot}225 + 0{\cdot}025\ (C - 4{\cdot}05)^2$ has been deduced.*



4.2. *Supplementary Experiments*

During the writing of this paper some incidental qualitative experiments were made.

(a) Using a direct-reading noise-factor test set[12] at 45 Mc/s, the capacitance detuning required for optimum noise factor has been measured on a large number of valves of various types. The experimental accuracy in determining the detuned conditions is very good with this equipment, and on all valves tested (except one sample* of CV408) the detuning required was significantly less than the space-charge capacitance.

*Or, rather one measurement.

(b) Measurements of the effect of stray wiring leads in transformation of admittance were made at 45 Mc/s. These showed that leads of $\frac{3}{4}$ in. could cause a 10 or 20% error in the conductance term with the typical values used at 45 Mc/s. In other words, the source conductance at the grid terminal might be in error by such an amount unless measured at that terminal.

(c) The effect of cathode-lead inductance on noise factor was checked at 45 Mc/s on a unit using a 6AK5 with cathode-lead inductance to give an input conductance of 500 micromhos. With a constant source conductance of 500 micromhos, the cathode lead gave about 5% increase in first-stage noise factor and a similar further increase in noise factor due to the second stage.

(d) Using the direct-reading unit, the effect of feedback on noise factor has been investigated with a capacitance-neutralized circuit. It was found that minimum noise factor was obtained when the neutralizing was detuned to compensate the grid–cathode detuning, although the change in noise factor from the value with normal neutralization was very small. With 'tuned' input circuit, the neutralized condition gives minimum noise factor.

When the neutralizing capacitance was detuned in either sense from this position leaving the input circuit tuning unchanged, the noise factor was degraded, perhaps by as much as 1 db before the first stage went into oscillation.

This increase in noise factor due to anode–grid feedback was not due to increase in second-stage noise.

5. Discussion of Results

If we express the induced grid noise as equivalent to a correlated part $\{(a + jb)i_a\}^{2x}$ where

TABLE 1

VALVE PARAMETERS

Valve Type	V_A (V)	I_A (mA)	g_m (μA/V)	R_g (kΩ)	R_{equ} (Ω) (at 45 Mc/s)
CV139	250	10	9460	9·7	372
6AK5	110	8	5930	4·8	632
*VX3052	150	10	7250	7·5	544

* This early development sample had thicker grid wires, etc., than the CV408 (A1714).

TABLE 2

'TUNED' NOISE FACTOR ANALYSIS

Valve Type	Freq. (Mc/s)	A	B (Ω)	D	$G_D + G_T$	G_N	Uncorrelated Term: $D - (G_D + G_T + G_N + R_{equ}b^2g_m^2)$	† $G_T + G_K$ (measured)	G_o
				mhos × 10^{-6}					
CV139	12	1·07	368	322	35	2	250	34	935
CV139	30	1·05	427	482	27	4	310	80	1060
CV139	45	1·05	366	513	— 83 —		270	92	1180
CV139	60	1·06	360	802	110	14	530	445	1490
6AK5	45	1·02	607	135	$G_D = 33$? (see text)	3		63	470
VX3052	45	1·05	567	401	72	3	305	77	770

†Much of the variation would be due to differences in cathode decoupling. G_K is the conductance due to cathode-lead inductance.

TABLE 3

'DETUNED' NOISE FACTOR ANALYSIS

Valve Type	Freq. (Mc/s)	G_A Source Conductance (mhos × 10^{-6})	n_o (E)	C_o (pF)	ΔC (measured) (pF)	F × 10^{24}	$\frac{FG_A}{\omega^2}$ (Ω)	Correlated Term: $C_o^2FG_A$ (mhos × 10^{-6})
CV139	12	77·5	4·77	3·1	3·1	·027	380	20
	12	410	1·98	3·4		·0064	465	30
CV139	30	320	2·33	2·8	3·3	·048	435	120
	30	539	2·00	3·0$_5$		·032	490	161
	30	1310	1·93	3·0		·011	410	130
CV139	45	248	2·5	2·1	2·9	·127	400	149
	45	920	1·81	2·3		·038	440	182
	45	2095	2·00	2·2		·015	390	152
	45	3390	2·38	1·9		·0079	326	111
CV139	60	335	3·2	1·5$_5$	3·2	·192	453	154
	60	834	2·31	1·6		·075	440	160
	60	2390	2·22	1·7		·025	422	163
6AK5	45	221	1·54	·7	1·0	·19	530	20·5
	45	897	1·66	·7		·047	530	21
VX3052	45	226	2·18	·8		·163	460	24
	45	330	2·05	·9	1·4	·123	510	31
	45	915	1·83	·9		·044	500	35
	45	1360	2·11	1·1		·023	390	36

ΔC = Measured increase in capacitance due to space-charge.

$i_a^{2\Sigma}$ is the anode noise current, and an uncorrelated term $t4kT_0G_T$, then we get (Fig. 5)

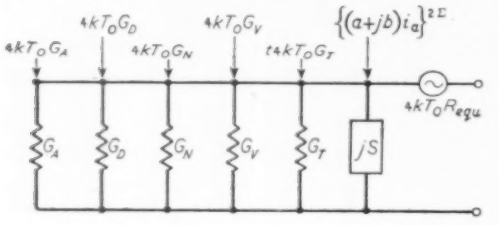


Fig. 5. *Equivalent noise circuit with induced grid noise split into correlated and uncorrelated parts.*

$$n = 1 + \frac{G_D + G_N + G_V + tG_T}{G_A} + \frac{R_{equ}}{G_A}$$
$$\{(G_A + G_D + G_N + G_V + G_T + ag_m)^2 + (S + bg_m)^2\} \quad .. \quad .. \quad .. \quad (7)$$

where G_A = source conductance
G_D = input coil conductance
G_V = valve input losses not due to electronic emission, and assumed at temperature T_0
G_T = induced grid noise equivalent conductance not correlated wth shot noise
t = noise temperature ratio of G_T
G_N = neutralizing-coil conductance
S = input circuit susceptance excluding any susceptance due to feedback
R_{equ} = shot noise equivalent resistance, but including small terms to allow for second-stage noise and noise due to G_N generated between anode and cathode.
g_m = mutual conductance.

On re-arranging the terms and equating with the experimental expressions, we get:—

When $S = 0$ ('Tuned' noise factor)

$$A = 1 + 2R_{equ}(ag_m + G_D + G_N + G_V + G_T)$$
$$B = R_{equ}$$
$$D = G_D + G_N + tG_T + G_V + R_{equ}\{(G_D + G_N + G_V + G_T + ag_m)^2 + b^2g_m^2\}$$
$$\approx G_D + G_N + tG_T + G_V + R_{equ}b^2g_m^2$$

Deduced from the experimental results for $A-1$

Hence $tG_T = D - (G_D + G_N + G_V + R_{equ}b^2g_m^2)$.

When $S = -bg_m$ ('Detuned' noise factor)

$$C_0 = \frac{-bg_m}{\omega}; \quad F = \frac{R_{equ}\omega^2}{G_A}$$

Therefore $R_{equ} = \frac{FG_A}{\omega^2}$; $C_0^2FG_A = R_{equ}b^2g_m^2$

The accuracy of the experimental fit is very good for the majority of the results. The CV139 results at 30 Mc/s were suspected of being faulty due to a noisy second stage, but a check-back on the unit used, after a long lapse of time, showed that this was not the cause of the discrepancy in the value of B at this frequency. In view of the discrepancy in the tuned noise factor results at 30 Mc/s (Table 2), the other results at 30 Mc/s should be ignored. At 60 Mc/s, the value obtained for A seems too small, and is too small with the 6AK5 at 45 Mc/s.

Some of the possible causes of discrepancies are:—

(1) The effect of G_N on R_{equ} will be to give a term which is correlated exactly with the other term due to G_N. This will give very small error.

(2) The effect of the second stage is not exactly equivalent to a constant increase in the shot noise because of the dependence of the first stage output impedance on the source impedance with the coil-neutralized circuit. This could cause no more than about 5 ohms variation in R_{equ}. It would have been better to measure the noise factor of the first stage only.

(3) The supplementary results show that there is a significant change in noise factor with large values of lead inductance. This causes some doubts about errors due to neglect of cathode-lead inductance at the highest frequency.

(4) There is probably some significant error in the measurement of G_A for small values at the higher frequencies.

(5) The coils and valve cold losses are assumed to generate thermal noise equivalent to 290°K. The coils were in fact close to ambient temperature, but the valve losses may be at a much higher temperature.

(6) Due to an oversight, the input losses were not separated, and at 45 Mc/s, although the same input coil was used for the three valves, there was no separate measurement on the 6AK5. A separate measurement was made of the coil, giving 33 micromhos, but this was not in situ and was probably lower than the value when connected in the circuit.

It is doubtful whether a more detailed analysis would be of practical value in view of the experimental limitations. In qualitative terms, the following points have been considered.

(7) The effect of cathode-lead inductance does not give feedback of noise sources between grid and earth, anode and cathode, and grid and cathode, in exactly the same ratio. See Fig. 6.

(8) The correlated part of induced grid noise is strictly equivalent to the transit-time effects of all electrons which induce noise at both grid and anode during their flight, and the

strict representation should include a generator between grid and anode. However, this grid–anode generator can be replaced by equivalent correlated generators, one between anode and cathode and the other between grid and cathode. The analysis has allowed for this, but not for the input admittance due to the correlated component.

Some features of the results are:—

(a) The large value of uncorrelated noise relative to the correlated term.

(b) C_0 is a function of frequency, and is nearly equal to the space-charge capacitance at low frequencies.

(c) That the correlated, uncorrelated and total induced grid noise components do not obey the theoretical law in relation to frequency ($\propto f^2$).

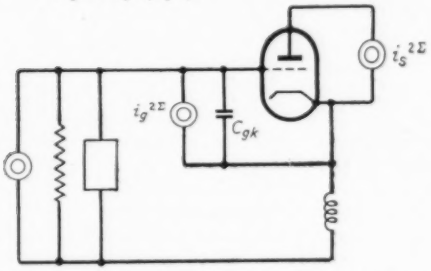


Fig. 6. *Cathode-lead inductance affects the feedback of noise sources differently, according to their positions.*

The results do not show striking agreement with the simple theory of correlation outlined, and for which R. L. Bell[7] has obtained close experimental agreement. The complexity of the theoretical expressions and the experimental work (together with the lack of detailed information about the experimental methods used by Bell) hardly permit much inference from the discrepancies noted.

It appears that the advantages to be obtained by detuning will be very small at frequencies much higher than those investigated, and cathode-lead inductance may well be part of the cause. On the other hand, reduction of frequency does not appear to give such a large reduction of induced grid noise as would be expected. This may be due to other grid–cathode losses. Some confirmation of the limited improvement at lower frequencies is given by other work. Even with valves giving 1·5 db noise factor at 45 Mc/s, a noise factor better than 1·0 db has not been achieved at 10 Mc/s, despite the use of very large low-loss coils.

6. Conclusion

The discrepancy between simple approximate theory and practice (neglecting correlation) noted elsewhere[4] seems to be due to the fact that the approximations made there were more serious than might be thought. Most probably, the value of source conductance giving optimum noise factor was greater than deduced from simple estimates. From the present results the fit with theoretical expressions is very good, having taken precautions to measure G_A accurately, but the constants for induced grid noise do not agree with theory.

There seems little hope of eliminating the effects of induced grid noise with conventional valves except by direct improvement of the valves. An easing of the problem of close spacings may perhaps be obtainable with the space-charge grid valve reported by Peterson[14] and tried by Neher[15] at 3,000 Mc/s. This valve gives a virtual emission of faster electrons, and a high mutual conductance.

A more detailed investigation of noise factor at the lower frequencies seems to be required. Although Wallmann, Macnee and Gasden[16] have reported a noise factor of $\frac{1}{4}$ db (1·06) at 6 Mc/s with a 6AK5, a more complete understanding of the reasons for the virtues of this valve (if substantiated) might point the way for further progress.

We close with a quotation from North[17] on the subject of transit-time effects on noise:—

"Although there is every need for an understanding of shot fluctuations at ultra-high frequencies, it comes not on the wings of the morning, ..."

7. Acknowledgment

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APPENDIX

Calculation of Noise Factor with Correlation

A. $$n = \frac{\text{Available Power}}{\text{Power available from source}}$$

If in Fig. 5, $G = G_A + G_D + G_N + G_F + G_T$.

Then voltage² due to correlated induced grid noise and shot noise per unit frequency interval

$$= \left[\frac{(a+jb)i_a}{G+jS} + \frac{i_a}{g_m}\right]^{2\Sigma}$$

$$= \left[\frac{\{g_m(a+jb)+(G+jS)\}\,i_a}{g_m(G+jS)}\right]^{2\Sigma}$$

$$= \frac{\{(ag_m+G)^2+(S+bg_m)^2\}\,i_a^{2\Sigma}}{g_m^2(G^2+S^2)}$$

Available power due to this term

$$= \frac{\{(ag_m+G)^2+(S+bg_m)^2\}\,i_a^{2\Sigma}}{4g_m^2G}$$

$$= \frac{\{(ag_m+G)^2+(S+bg_m)^2\}\,4kT_oR_{equ}}{4G}$$

But, since available power from $G_A = \frac{4kT_oG_A}{G}$, the term due to shot noise in the expression for noise factor is

$$\frac{\{(ag_m + G)^2 + (S + bg_m)^2\} R_{eqv}}{G_A}$$

The other terms are obtained very easily.

B. The input admittance due to cathode-lead inductance is given by:—

$$Y_K = \frac{\omega^2 L_K C_{gk} g_m + j\omega C_{gk}(1 - \omega^2 L_k C_{gk})}{(1 - \omega^2 L_k C_{gk})^2 + \omega^2 L_k^2 g_m^2}$$

Inserting approximate values for g_m, C_{gk}, L_k gives values for change of input capacitance of less than 2% at the frequencies used. By calculating the value of L_k required to give an input conductance of 500 micromhos at 45 Mc/s with a 6AK5, it is deduced that the change of input capacitance is about -25% of C_{gk}, or taking $C_{gk} = 3$ pF, about $-0{\cdot}75$ pF. This is closely equal to the capacitance detuning required for optimum noise factor.

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CORRESPONDENCE

Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

Matching Discontinuities in Waveguides

SIR,—In his paper in the October 1953 issue of *Wireless Engineer* (p. 243), Mr. Parr states "Further, it is usually impossible to reduce the standing-wave ratio of an unmatched component to zero (i.e., to match the component) if imperfect measuring equipment is used."

I would point out that, in the matched condition, the standing-wave ratio is unity, not zero. The latter condition corresponds to perfect reflection.

PREM SWARUP.

Dept. of Physics,
University of Allahabad, India.
5th December 1953.

[The author writes to say in reply that he agrees with Mr. Swarup's comment and that he inadvertently wrote 'standing-wave ratio' instead of 'reflection coefficient'. —Ed.]

NEW YEAR HONOURS

In the New Year Honours List, W. G. Radley (Engineer-in-Chief, G.P.O.) received a knighthood and N. C. Robertson (Deputy Managing Director, E. K. Cole; until recently, Director-General of Electronics Production, Ministry of Supply) becomes a Companion of the Order of St. Michael and St. George, while Commander S. S. C. Mitchell, R.N.(Ret.) (Controller of Guided Weapons and Electronics, Ministry of Supply), becomes a Knight-Commander of the British Empire.

G. Darnley Smith (Managing Director, Bush Radio), and G. J. S. Little (Assistant Engineer-in-Chief, G.P.O.) were appointed Commanders of the Order of the British Empire.

T. H. Baines (Deputy Director of Radio Equipment [Production] Admiralty), A. S. Mitson (Assistant Director, Electronic Production, Ministry of Supply) and M. J. L. Pulling (Senior Superintendent Engineer, Television, B.B.C.) become Officers of the Order of the British Empire.

O. H. Barron (Engineer, Planning & Installation Department, B.B.C.), W. H. F. Griffiths (Chief Engineer, H. W. Sullivan), H. J. Harbour (Test Controller, E. K. Cole) were appointed Members of the Order of the British Empire.

I.E.E.

The Council of the Institution of Electrical Engineers has made the 32nd award of the Faraday Medal to Isaac Shoenberg "for his distinguished work in electrical engineering, in particular the outstanding contributions which he has made to the development of high-definition television in this country". Mr. Shoenberg joined E.M.I., Ltd., in 1931 as Director of Research and was the leader of the team of scientists which developed the television system adopted by the B.B.C. in 1936. Born in Russia in 1880, he came to this country in 1914 to take up an appointment with Marconi's Wireless Telegraph Company.

NEW APPOINTMENTS

A. H. Mumford, O.B.E., B.Sc.(Eng.), M.I.E.E., has been appointed Deputy Engineer-in-Chief of the G.P.O. Engineering Department in succession to H. Faulkner, C.M.G., B.Sc., M.I.E.E., F.I.R.E., who is retiring from the Post Office after 40 years' service.

Mr. Faulkner has been appointed a Director of the Telecommunication Engineering & Manufacturing Association.

EXHIBITIONS

The 38th Annual Exhibition of the Physical Society will be held at the Imperial College of Science and Technology, Imperial Institute Road, London, S.W.7, from Thursday, 8th April to Tuesday, 13th April. Applications for tickets should be made to the Secretary-Editor, 1 Lowther Gardens, Prince Consort Road, London, S.W.7.

The exhibition of components, organized by the Radio & Electronic Component Manufacturers' Federation, will be held at Grosvenor House, Park Lane, London, W.1, from Tuesday 6th April to Thursday 8th April. Applications for tickets should be made to the Federation at 22 Surrey Street, Strand, London, W.C.2.

NORMALIZED IMPEDANCE AND REFLECTION COEFFICIENT

Graphical Representation of their Relationship

By P. A. Lindsay, Ph.D., D.I.C., A.C.G.I.

(Communication from the Staff of the Research Laboratories of The General Electric Co., Ltd., Wembley, England)

IN the circuit analysis of most transmission systems, such as cables, waveguides, aerials, etc., it is often desirable to express the circuit characteristics of terminating or load impedances $\mathbf{Z}$ in terms of the voltage coefficient of reflection $\boldsymbol{\rho}$. The relationship between $\mathbf{Z}$ and $\boldsymbol{\rho}$ can be derived quite simply from the expression for the impedance $\mathbf{Z}$ at the end of a transmission line of length l and propagation constant $\mathbf{P}$,

$$\mathbf{Z} = \mathbf{Z}_0 \frac{\mathbf{A} \exp(-\mathbf{P}l) + \mathbf{B} \exp \mathbf{P}l}{\mathbf{A} \exp(-\mathbf{P}l) - \mathbf{B} \exp \mathbf{P}l}$$

where $\mathbf{Z}_0$ is the characteristic impedance of the line (see for example Ref. 1, p. 154 et seq.).

If we define $\boldsymbol{\rho}$ as the ratio $\{\mathbf{B} \exp \mathbf{P}l\}/\{\mathbf{A} \exp(-\mathbf{P}l)\}$ of the reflected to the incident voltage at the end of a transmission line, we get from the equation for $\mathbf{Z}$ the following expressions for the relationship between $\mathbf{Z}$ and $\boldsymbol{\rho}$:

$$\mathbf{z} = \frac{1+\boldsymbol{\rho}}{1-\boldsymbol{\rho}} \quad (1) \qquad \mathbf{y} = \frac{1-\boldsymbol{\rho}}{1+\boldsymbol{\rho}} \quad (2)$$

$$\boldsymbol{\rho} = \frac{\mathbf{z}-1}{\mathbf{z}+1} \quad (3) \qquad \boldsymbol{\rho} = \frac{1-\mathbf{y}}{1+\mathbf{y}} \quad (4)$$

where $\mathbf{z} = \dfrac{\mathbf{Z}}{\mathbf{Z}_0} = r + jx$ = normalized load impedance

$\mathbf{y} = \dfrac{\mathbf{Y}}{\mathbf{Y}_0} = g + jb$ = normalized load admittance

$\boldsymbol{\rho} = \rho \exp j\phi$ = voltage coefficient of reflection.

$\mathbf{Z}_0$ and $\mathbf{Y}_0$ being the characteristic impedance and admittance of the line.

Expressions (1) – (4) refer to complex quantities and belong to the bi-linear group of conformal transformations. Ideally they require four dimensions for full representation of the variables r, x, ρ and ϕ (or g, b, ρ and ϕ) along four orthogonal axes. Fortunately, expressions (1) – (4) can also be represented in two dimensions, although with some loss of clarity, either by perspective drawings of a pair of three-dimensional surfaces, one pair for each equation, or by a simultaneous projection of the contours of each pair of surfaces on to a plane. The contour representation of the functions $\boldsymbol{\rho}$, $\mathbf{z}$ and $\mathbf{y}$ was originally suggested by Smith,[2] in his paper on circle diagrams. This method, although excellent for computational purposes, does not appeal very strongly to our visual imagination. For a beginner it is of little help in tracing the intricate relationships which exist between the six quantities.* Yet in spite of some excellent examples from other fields[3] no perspective drawings of the surfaces representing the important functions $\boldsymbol{\rho}$, $\mathbf{z}$ and $\mathbf{y}$ are available in literature. It seems to the author that this gap should be filled in order to assist those who prefer pictorial to purely algebraic representation of the functions (1) – (4).

The first step in representing expressions (1) – (4) graphically is to split $\boldsymbol{\rho}$, $\mathbf{z}$ and $\mathbf{y}$ into their real and imaginary parts. From equation (1) we get, after multiplying numerator and denominator by $1 - \rho \exp(-j\phi)$,

$$\begin{aligned}\mathbf{z} &= \frac{1+\boldsymbol{\rho}}{1-\boldsymbol{\rho}} \\ &= \frac{1+\rho\exp(j\phi)}{1-\rho\exp(j\phi)} \cdot \frac{1-\rho\exp(-j\phi)}{1-\rho\exp(-j\phi)} \\ &= \frac{1-\rho^2+j2\rho\sin\phi}{1+\rho^2-2\rho\cos\phi}\end{aligned}$$

or remembering that $\mathbf{z} = r + jx$

$$r = \frac{1-\rho^2}{1+\rho^2-2\rho\cos\phi} \quad \ldots \quad (5)$$

$$x = \frac{2\rho\sin\phi}{1+\rho^2-2\rho\cos\phi} \quad \ldots \quad (6)$$

Similarly from equation (2) and from the relation $\mathbf{y} = g + jb$ we get:

$$g = \frac{1-\rho^2}{1+\rho^2+2\rho\cos\phi} \quad \ldots \quad (7)$$

$$b = \frac{-2\rho\sin\phi}{1+\rho^2+2\rho\cos\phi} \quad \ldots \quad (8)$$

In all four cases we get $\mathbf{z}$ or $\mathbf{y}$ in terms of the polar components of $\boldsymbol{\rho}$. For the reverse transformation we rationalize the right-hand side of equation (3) and get:

$$\boldsymbol{\rho} = \frac{\mathbf{z}-1}{\mathbf{z}+1} = \frac{r-1+jx}{r+1+jx} = \frac{r^2+x^2-1+j2x}{(r+1)^2+x^2}$$

which leads to

MS accepted by the Editor, March 1953

* It was brought to my notice by Dr. Bloch that the approach to circle diagrams suggested in his paper[4] provides a form of graphical representation which could also appeal to our visual imagination.

$$\rho = \sqrt{\frac{(r-1)^2 + x^2}{(r+1)^2 + x^2}} \quad .. \quad .. \quad .. \quad (9)$$

$$\phi = \tan^{-1}\frac{2x}{r^2 + x^2 - 1} \quad .. \quad .. \quad .. \quad (10)$$

By a similar process we get from equation (4):

$$\rho = \sqrt{\frac{(1-g)^2 + b^2}{(1+g)^2 + b^2}} \quad .. \quad .. \quad .. \quad (11)$$

$$\phi = \tan^{-1}\frac{2b}{g^2 + b^2 - 1} \quad .. \quad .. \quad .. \quad (12)$$

It is worth noting at this point that because of the simple inverse relationship between **z** and **y** the number of different representations of the functions (5) – (12) is halved and many surfaces can be obtained merely by a shift or rotation of the right set of co-ordinate axes, as shown in Figs. 1–4.

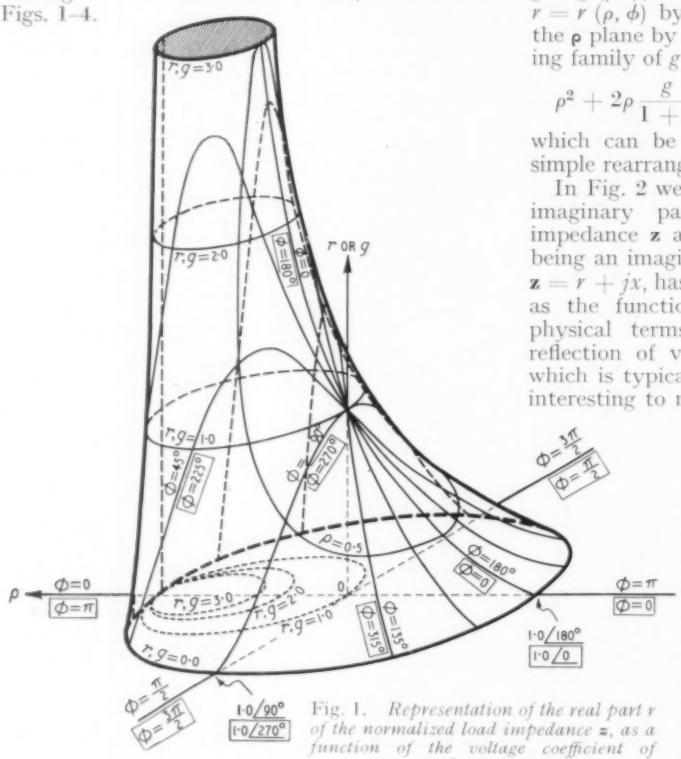


Fig. 1. *Representation of the real part r of the normalized load impedance z, as a function of the voltage coefficient of reflection* ρ.

Let us consider now each of the surfaces in turn. We have in Fig. 1 the representation of r as a function of ρ and phase angle ϕ, where, for convenience ρ and ϕ are expressed in polar co-ordinates. For reasons which will become clear later, the function r is considered here for $\rho \leqslant 1$ only. The function r has a simple pole at $\rho = 1$, $\phi = 0$, corresponding to an open circuit, causing a total reflection at the end of the line. The thin lines on the surface $r = r\ (\rho, \phi)$ represent cross-sections along the ϕ = constant planes and along one ρ = constant cylinder. The horizontal contours, together with their dotted projections on the ρ plane, correspond to the r = constant curves. They can be derived from equation (5) by treating r as a parameter of a family of curves. Thus rearranging terms in equation (5) we get:

$$\rho^2 - 2\rho\frac{r}{1+r}\cos\phi = \frac{1-r}{1+r} \quad .. \quad .. \quad (13)$$

which is the equation of the r = const. family of circles in the polar form of the circle diagram. In the figure the values of the constants are marked on the corresponding contours. Further, it is shown by the boxed figures that the related $g = g\ (\rho, \phi)$ surface can be obtained from $r = r\ (\rho, \phi)$ by rotating the co-ordinate axes in the ρ plane by an angle of 180°. The corresponding family of g = const. curves is given by

$$\rho^2 + 2\rho\frac{g}{1+g}\cos\phi = \frac{1-g}{1+g} \quad .. \quad .. \quad (14)$$

which can be derived from equation (7) by a simple rearrangement of terms.

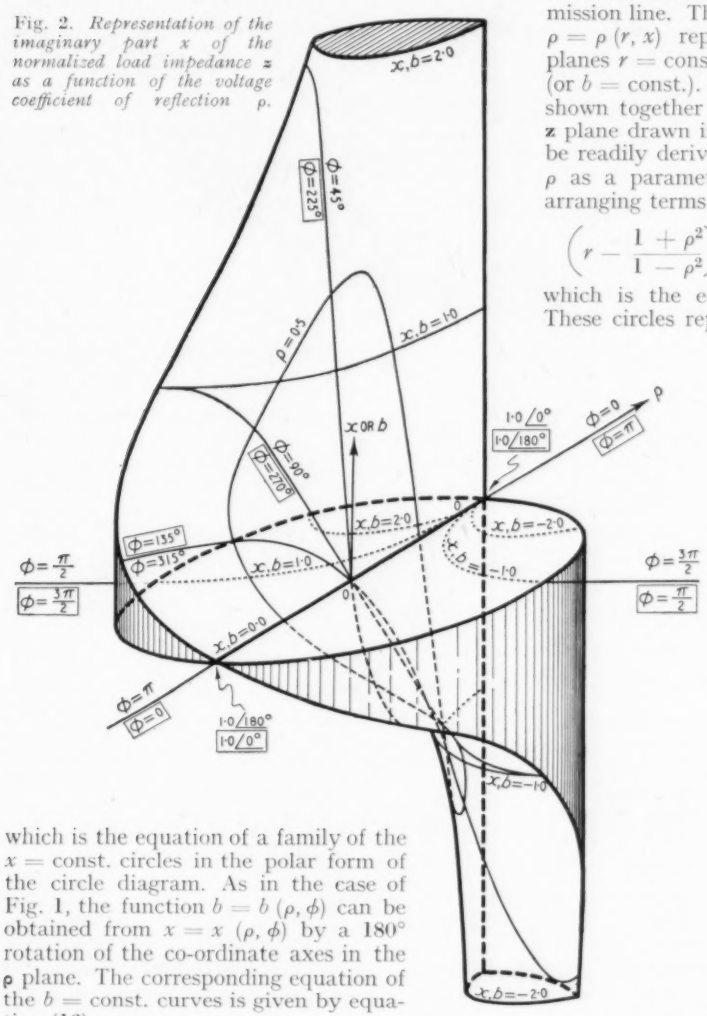
In Fig. 2 we have the surface representing the imaginary part x of the normalized load impedance **z** as a function of ρ. This function being an imaginary part of the complex variable $\mathbf{z} = r + jx$, has a discontinuity at the same point as the function r; i.e., at $\rho = 1$, $\phi = 0$. In physical terms this point represents a total reflection of voltage without change of phase, which is typical for an open-circuited line. It is interesting to note that, unlike r, the function x is different from zero for all points on the circumference of the unit circle $\rho = 1$ (except for $\phi = \pi$). The difference in the behaviour of the functions r and x is due to the fact that the magnitude of ρ (but not phase) remains constant and equal to one, whatever the value of the load, as long as the normalized load is purely reactive. However, as soon as the normalized load exhibits resistive components some power loss is bound to occur, and some reduction in the magnitude of ρ must follow, as can be shown in Fig. 1. We can also see from Fig. 2 that the function x changes sign depending on the phase of ρ. It is positive (inductive) for $0 < \phi < \pi$ and negative (capacitive) for $\pi < \phi < 2\pi$. For $\phi = 0$ and $\phi = \pi$ we have conditions of purely resistive normalized loads correspondingly greater and less than one.

The ρ = const. ϕ = const. and x = const. cross-sections are shown on the figure in thin

lines, and the projections of the $x = \text{const.}$ contours on to the ρ plane are dotted. The equation of the family of $x = \text{const.}$ curves can be derived from (6) by treating x as a constant parameter. Thus rearranging terms we get:

$$\rho^2 - 2\rho\left(\cos\phi + \frac{1}{x}\sin\phi\right) = -1 \quad .. \quad (15)$$

Fig. 2. *Representation of the imaginary part x of the normalized load impedance $\mathbf{z}$ as a function of the voltage coefficient of reflection ρ.*



which is the equation of a family of the $x = \text{const.}$ circles in the polar form of the circle diagram. As in the case of Fig. 1, the function $b = b(\rho, \phi)$ can be obtained from $x = x(\rho, \phi)$ by a 180° rotation of the co-ordinate axes in the ρ plane. The corresponding equation of the $b = \text{const.}$ curves is given by equation (16),

$$\rho^2 - 2\rho\left(-\cos\phi - \frac{1}{b}\sin\phi\right) = -1 \quad .. \quad (16)$$

which can be derived from equation (8) by a simple rearrangement of terms.

Fig. 3 gives us the magnitude ρ of the voltage coefficient of reflection $\boldsymbol{\rho}$ as a function of the real and imaginary parts of the normalized load impedance $\mathbf{z}$. This function is zero only at $r = 1$, $x = 0$, which point corresponds to a matched or infinitely-long transmission line. For all other points of the $\mathbf{z}$ plane the function is $0 < \rho \leqslant 1$; i.e., some reflection must occur whenever $\mathbf{z} \neq 1$, or whenever the terminating load differs from the characteristic impedance of the transmission line. The thin lines drawn on the surface $\rho = \rho(r, x)$ represent the cross-section by the planes $r = \text{const.}$ (or $g = \text{const.}$) and $x = \text{const.}$ (or $b = \text{const.}$). The contours $\rho = \text{const.}$ are also shown together with their projections on to the $\mathbf{z}$ plane drawn in dotted lines. The contours can be readily derived from equation (9) by treating ρ as a parameter of a family of curves. Rearranging terms we get:

$$\left(r - \frac{1+\rho^2}{1-\rho^2}\right)^2 + x^2 = \left(\frac{2\rho}{1-\rho^2}\right)^2 \quad .. \quad (17)$$

which is the equation of a family of circles. These circles represent the $\rho = \text{const.}$ curves in the Cartesian form of the circle diagram. The surface shown in Fig. 3 represents both the $\rho = \rho(r, x)$ and $\rho = \rho(g, b)$ functions and similarly to equation (17) we can derive from equation (11) the following expression for the $\rho = \text{const.}$ circles in the $\mathbf{y}$ plane.

$$\left(g - \frac{1+\rho^2}{1-\rho^2}\right)^2 + b^2 = \left(\frac{2\rho}{1-\rho^2}\right)^2 \quad .. \quad (18)$$

The surface shown in Fig. 4 is more complicated than any of the others. It represents one branch of the essentially multi-valued function $\phi = \phi(r, x)$. This function becomes indeterminate at the point $r = 1$, $x = 0$, which point corresponds to a matched transmission line. It is obvious that at this point, in view of $\boldsymbol{\rho} = 0$, no meaning can be attached to the phase angle ϕ. For other values of $\mathbf{z}$, the chosen branch of the function ϕ lies between 0 and 2π. As previously, the thin lines represent the cross-sections of the surface with the $r = \text{const.}$ (or $g = \text{const.}$) and $x = \text{const.}$ (or $b = \text{const.}$) planes. The contours $\phi = \text{const.}$ are also shown together with their projections in dotted lines on to the $\mathbf{z}$ plane. The expression for the contours can be derived from equation (10) by treating ϕ as a parameter of a family of curves.

Rearranging terms we get from equation (10):

$$r^2 + (x - \cot\phi)^2 = \operatorname{cosec}^2\phi \quad .. \quad .. \quad (19)$$

which is the equation of a family of circles representing the $\phi = \text{const.}$ curves in the Cartesian form of the circle diagram. The related function $\phi = \phi\ (g, b)$ can be obtained from $\phi = \phi(r, x)$ by shifting the origin by π along the ϕ axis, as shown boxed. This shift is equivalent to the 180° rotation of axes also shown boxed in Figs. 1 and 2. The family of $\phi = \text{const.}$ contours can now be derived from equation (12) in terms of g and b. Rearrangement of terms in equation (12) gives us

$$g^2 + (b - \cot\phi)^2 = \operatorname{cosec}^2\phi \quad .. \quad .. \quad (20)$$

which is an equation of the family of circles representing the $\phi = \text{const.}$ curves in the Cartesian form of the circle diagram.

It should be added that in some cases it is necessary to consider values of r and $g < 0$ and $\rho > 1$ even for passive networks. The reason for this can be traced back to our definition of the variables $\mathbf{z}$ and $\mathbf{y}$. Thus for instance in the case of $\mathbf{z}$ we could easily have $\mathbf{Z} = Z/\underline{-90°}$ and $\mathbf{Z_0} = Z_0/\underline{10°}$ giving us $\mathbf{z} = \mathbf{Z}/\mathbf{Z_0} = Z/Z_0/\underline{-100°}$ which has its real part negative and leads to the values of $\rho > 1$ (in practice the reactive component of $\mathbf{Z_0}$ is normally fairly small causing ρ to be only slightly greater than one). In the case of negative r and g and ρ greater than one Figs. 1-4 still can be used, but only if we introduce an auxiliary variable

$$\boldsymbol{\rho}' = \frac{1}{\boldsymbol{\rho}^*} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (21)$$

where $\boldsymbol{\rho}^* = \rho \exp(-j\phi)$ and is the complex conjugate of the voltage coefficient of reflection $\boldsymbol{\rho}$. Equation (21) defines a conformal transformation which causes the whole of the $\boldsymbol{\rho}$ plane outside the unit circle $\rho = 1$ to be represented by the points inside the circle. The transformation is so arranged that the magnitude of $\boldsymbol{\rho}$ is changed, but its phase angle ϕ remains the same. Expressing

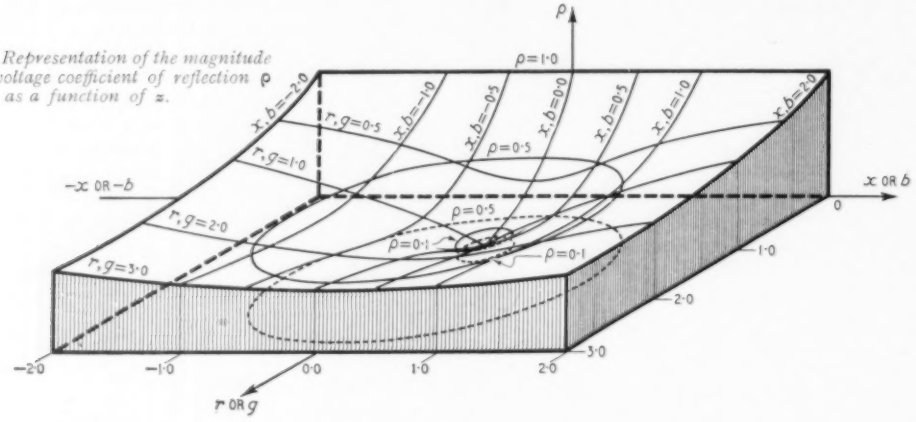


Fig. 3. *Representation of the magnitude ρ of the voltage coefficient of reflection ρ as a function of z.*

equation (21) in terms of magnitude and phase we find that

$$\rho' = \frac{1}{\rho} \quad .. \quad (22) \quad \text{and} \quad \phi' = \phi \quad .. \quad (23)$$

as desired.

From equations (22) and (23) we can see that equation (21) can be also expressed in the form

$$\boldsymbol{\rho}'^* = \frac{1}{\boldsymbol{\rho}} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (24)$$

Substituting equations (21) and (24) in equation (1) we get:

$$\mathbf{z} = \frac{\boldsymbol{\rho}'^* + 1}{\boldsymbol{\rho}'^* - 1} = \frac{\rho'^2 - 1 + j2\rho' \sin\phi'}{\rho'^2 + 1 - 2\rho' \cos\phi} \quad .. \quad .. \quad (25)$$

or separating the real and imaginary parts

$$r = -\frac{1 - \rho'^2}{1 + \rho'^2 - 2\rho' \cos\phi'} \quad .. \quad .. \quad (26)$$

$$x = \frac{2\rho' \sin\phi'}{1 + \rho'^2 - 2\rho' \cos\phi'} \quad .. \quad .. \quad (27)$$

Similarly we get for the components g and b of the normalized admittance $\mathbf{y}$ in terms of the new variable $\boldsymbol{\rho}'$

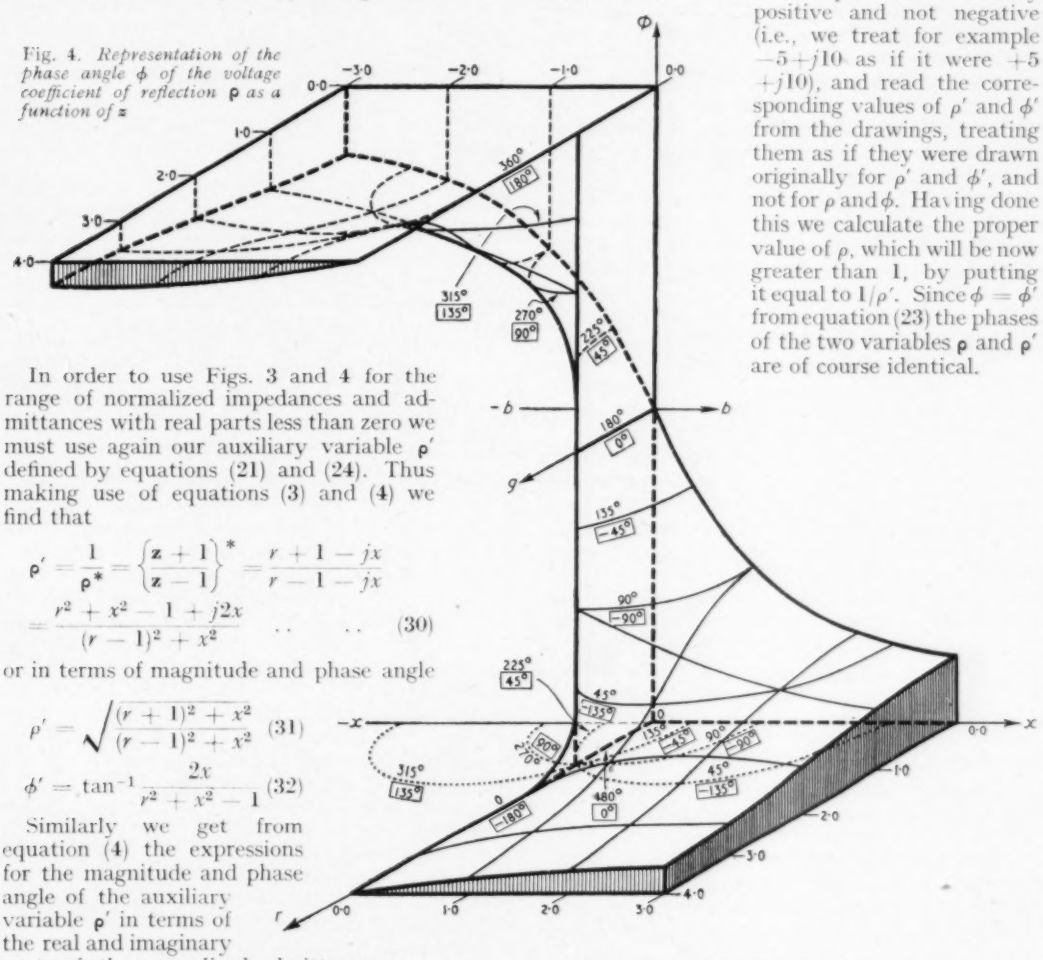
$$g = -\frac{1 - \rho'^2}{1 + \rho'^2 + 2\rho' \cos\phi'} \quad .. \quad .. \quad (28)$$

$$b = -\frac{2\rho' \sin\phi'}{1 + \rho'^2 + 2\rho' \cos\phi'} \quad .. \quad .. \quad (29)$$

Comparing expressions for $\mathbf{z}$ and $\mathbf{y}$ in terms of $\boldsymbol{\rho}$ and $\boldsymbol{\rho}'$ we find that they are identical, except

for the sign of r and g. Thus if we wish to use Figs. 1 and 2 for finding the values of **z** or **y** corresponding to a given value of ρ, where $\rho > 1$, we form first of all a new variable ρ', by putting $\rho' = 1/\rho$ and $\phi' = \phi$. Having done this we use Fig. 1 as if it were drawn originally for ρ' and not ρ, and read from it the value of r or g, adding at the same time a negative sign in front of it. Similarly we treat Fig. 2 as if it were drawn for ρ' and not ρ, and read from it directly the value of x or b, without any change of sign at all.

Fig. 4. *Representation of the phase angle ϕ of the voltage coefficient of reflection ρ as a function of* **z**



In order to use Figs. 3 and 4 for the range of normalized impedances and admittances with real parts less than zero we must use again our auxiliary variable ρ' defined by equations (21) and (24). Thus making use of equations (3) and (4) we find that

$$\rho' = \frac{1}{\rho^*} = \left\{\frac{\mathbf{z}+1}{\mathbf{z}-1}\right\}^* = \frac{r+1-jx}{r-1-jx}$$

$$= \frac{r^2+x^2-1+j2x}{(r-1)^2+x^2} \quad .. \quad .. \quad (30)$$

or in terms of magnitude and phase angle

$$\rho' = \sqrt{\frac{(r+1)^2+x^2}{(r-1)^2+x^2}} \quad (31)$$

$$\phi' = \tan^{-1}\frac{2x}{r^2+x^2-1} \quad (32)$$

Similarly we get from equation (4) the expressions for the magnitude and phase angle of the auxiliary variable ρ' in terms of the real and imaginary parts of the normalized admittance **y**

$$\rho' = \sqrt{\frac{(1+g)^2+b^2}{(1-g)^2+b^2}} \quad .. \quad .. \quad .. \quad (33)$$

$$\phi' = \tan^{-1}\frac{2b}{g^2+b^2-1} \quad .. \quad .. \quad .. \quad (34)$$

From equations (9), (11) and (31), (33) we find that functions ρ and ρ' satisfy the following symmetry relationship

$$\rho'(-r, x) = \rho(r, x) \quad .. \quad .. \quad .. \quad (35)$$

$$\rho'(-g, b) = \rho(g, b) \quad .. \quad .. \quad .. \quad (36)$$

the functions ϕ and ϕ' being of course identical. Thus if we wish to use Figs. 3 and 4 for finding the value of the voltage coefficient of reflection ρ for a given value of **z** or **y**, where **z** and **y** have negative real parts, we treat the variables **z** and **y** first of all as if their real parts were actually positive and not negative (i.e., we treat for example $-5+j10$ as if it were $+5+j10$), and read the corresponding values of ρ' and ϕ' from the drawings, treating them as if they were drawn originally for ρ' and ϕ', and not for ρ and ϕ. Having done this we calculate the proper value of ρ, which will be now greater than 1, by putting it equal to $1/\rho'$. Since $\phi = \phi'$ from equation (23) the phases of the two variables ρ and ρ' are of course identical.

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PHYSICAL BASIS OF THERMAL NOISE

By D. A. Bell, M.A., B.Sc., Ph.D., M.I.E.E.

ALTHOUGH very convenient, Nyquist's thermodynamic derivation of the formula for Johnson noise in a resistance under conditions of thermal equilibrium has in some ways been a misfortune for engineers: it has led so many people to ignore entirely the *mechanism* of thermal noise and to make the mistake of thinking that there is an inevitable connection between 'thermal noise' and 'thermal equilibrium'. The difficulties which have arisen in the application of the second law of thermodynamics to gases when the scale of observation is such as to show the statistical fluctuation, might perhaps be a warning against the care-free application of equilibrium arguments; and in any case there must always be a *mechanism* by means of which the equilibrium is established, and in metallic conductors the mechanism is very well established.

The Analogy with Brownian Motion

The experimental observations of Johnson[1] and the theoretical work of Nyquist[2] were published in 1928, but it had been suggested by de Haas Lorentz[3] as early as 1913 that there should be 'Brownian motion of electricity' in electric circuits. It must be noted, however, that the theory of Brownian motion, as usually expressed, describes a *total mean-square displacement*, whereas electrical engineering work from Nyquist onwards has been concerned with the *components of the disturbance in a narrow frequency band*. Thus there appears at first sight to be no relation between the formula of de Haas Lorentz,*

$$< i^2 > = X^2/2RL \quad . . \quad . . \quad . . \quad (1)$$

and that of Nyquist

$$< V^2_{df} > = 4RkT\,df \quad . . \quad . . \quad (2)$$

(The de Haas Lorentz formula relates to a series L, R circuit, and X is the total electromotive force equivalent to the results of all the collisions suffered by the conduction electrons.)

But these two formulae refer in fact to different quantities and are related to each other in the same way as the well-known formulae for total fluctuation due to pure shot noise in a CR circuit,

$$< V^2_{Total} > = \frac{2ie}{RC} \quad . . \quad . . \quad (3)$$

and the spectrum of the equivalent fluctuation current,

$$< I^2_{df} > = 2iedf \quad . . \quad . . \quad . . \quad (4)$$

Since the generalized form of (2) is

$$< V^2_{df} > = 4kT\mathrm{Re.}(Z)df \quad . . \quad . . \quad (5)$$

where Re. (Z)stands for the real or resistive part of Z, one can deduce the total thermal fluctuation in a CR circuit in the form

$$< V^2_{Total} > = 4kT \int_{f=0}^{f=\infty} \mathrm{Re.}\ (Z)df \quad . . \quad (5a)$$

and Moullin and Ellis[4] showed that this led to the equation

$$\tfrac{1}{2}C < V^2_{Total} > = \tfrac{1}{2}kT \quad . . \quad . . \quad (6)$$

The mean-square energy (over all frequencies) associated with the capacitance is thus the 'equipartition' value predicted by classical thermodynamic theory for any system with one degree of freedom. By taking the current-fluctuation version of Nyquist's formula,

$$< I^2_{df} > = (4kT/R)df \quad . . \quad . . \quad (7)$$

and working in admittances, one can arrive at a corresponding formula

$$\tfrac{1}{2}L < I^2_{Total} > = \tfrac{1}{2}kT \quad . . \quad . . \quad (8)$$

The results are readily obtained by direct integration for simple circuits consisting of one resistance and one reactance. For more complex circuits, however, one can obtain the corresponding general results by the technique which has been developed by Bode[5]: contour integration is used as a means of evaluating infinite integrals such as (5a) without using the values of the integrand at all real frequencies, as would be implied in simple integration. The method is readily applied to a passive two-terminal impedance (as distinct from a four-terminal or transfer impedance) of the type which Bode specified as a minimum-phase impedance; i.e., one free from singularities in the right-hand half of the 'p' (or 'complex frequency') plane and free from poles on the axis of real frequency. The first half of the condition requires that (a) any poles shall correspond to positive damping, thus excluding self-oscillating circuits and (b) the impedance shall be a single-valued function of frequency, a condition which may not be satisfied by the equations of distributed systems.

The process is detailed in Chapter 13 of Bode's book, and one of his results is as follows:

$$\int_0^\infty (A - A_\infty)\, d\omega = -\frac{\pi}{2} B_\infty \quad . . \quad . . \quad (9)$$

* The pointed brackets < > are used to denote *mean*-squared quantities.

MS accepted by the Editor, May 1953

with the notation that the impedance is $Z = A + jB$, where both A and B are functions of ω and, moreover, that Z is represented by a function which is analytic both at zero and infinite frequency. Then Z is to be expanded at infinite frequency in a series of inverse powers of ω:

$$Z = A_\infty + j\,B_\infty/\omega + A_1/\omega^2 + j\,B_1/\omega^3$$

The integral relation which has been quoted above means that if the real part of the impedance vanishes at infinite frequency ($A_\infty \to 0$) the infinite integral of A, the resistive part of the impedance, is equal to $-(\pi/2)\,B_\infty$. In general, a shunt capacitance will become dominant at infinite frequency, so that $j\,B_\infty/\omega$ will be $1/j\omega C_s$; i.e., $B_\infty = -1/C_s$. Under these conditions

$\int_0^\infty \text{Re. } (Z)\ d\omega = \pi/2C_s$ and the corresponding integral in *frequency* (instead of ω) will be $1/4C_s$. Then it follows that

$$< V^2_{Total} > =$$

$$\int_0^\infty < V^2_{df} > df = 4kT \int_0^\infty \text{Re. } (Z)df = kT/C_s \quad (10)$$

which demonstrates for the general case the result proved by Moullin and Ellis for particular cases, namely $\frac{1}{2}\,C < V^2\ _{Total} > = \frac{1}{2}kT$.

If the circuit becomes inductive at infinite frequency—for example a straight wire which would be almost a short-circuit at zero frequency—it can be treated in terms of admittance,

$$< I^2_{Total} > = 4kT \int_0^\infty \text{Re. } (Y)df = kT/L \quad (11)$$

which is equivalent to $\frac{1}{2}L < I^2_{Total} > = \frac{1}{2}kT$. This is the result which was obtained by Brillouin[6] from a consideration of the detailed behaviour of the conduction electrons in a metal.

'Degrees of Freedom' of a Circuit.

Two features of these conclusions deserve further attention. It might at first be thought that if the impedance under consideration consisted of a complex structure, perhaps with several separately identifiable resonant circuits, it could be considered to have several degrees of freedom and would therefore have a mean energy of several times the equipartition value $\frac{1}{2}kT$. The answer to this is that so long as the circuit is regarded as a two-terminal impedance, its state is fully specified when *the potential difference between the two terminals* is given at all times: it is therefore a system having only *one* degree of freedom when viewed from its terminals. The second question, which was raised by Moullin and Ellis,[4] is how the individual conduction electrons combine together to form a coherent whole which we call a 'circuit' having one degree of freedom. It is by no means certain that the existence of this coherence need be assumed since the mean-square voltage $<V^2>$ at the terminals may be merely the statistical resultant of numerous independent contributions. But in the work already cited Brillouin equated the 'inductive energy' $\frac{1}{2}Li^2$ of a circuit carrying steady current i to the kinetic energy of the conduction electrons in the circuit. It turned out that for a conductor of any appreciable size the 'mechanical' kinetic energy $\frac{1}{2}mv^2$ corresponding to the rest mass m of the electron and its drift velocity v was negligible in comparison with the electrical energy associated with the drift velocity of each electron in the field due to the common motion of all the other electrons, and the latter therefore represents the inductive energy. It is thus the common magnetic field associated with the common drift velocity of the electrons which provides some measure of 'coherence'.

Mechanism in a Metallic Conductor

This relationship between total fluctuation and spectrum intensity has been stressed because historically the first thorough theoretical analysis was published by Brillouin[6] in 1937 in the form

$$< I^2 > = kT/L$$

which corresponds to our formula (8) for total fluctuation. A simple quantitative model of the mechanism was published by the author[7] in 1938, and a closely similar model was used by Weisskopf[8] in 1943. The simple model will be summarized here because it can be used to illustrate the fact that 'resistance' and 'temperature' are really secondary parameters which may, in suitable cases, be used to specify the magnitude of thermal noise arising from the random motion of individual charges. Let an electron, with charge e, travel with velocity u for time τ. The product of charge by velocity gives not a current but a 'current-element' of the form $i\delta l$, and the electron-flight just proposed produces a pulse of current-element defined by

$$\int_0^\tau (i\delta l)dt = eu\tau \quad .. \quad .. \quad .. \quad .. \quad (12)$$

This pulse may be analysed into a spectrum of frequencies, and if N be the number of electrons present in the conductor it is found that their spectrum is represented by

$$< I^2_{df} > = [2N\,e^2u^2\tau/(\delta l)^2]df \quad .. \quad .. \quad (13)$$

But if the resistance of the conductor is deduced from the number of electrons present and their mobility,* the latter will be found to be $e\tau/2m$ and the resistance of the conductor is

* Mobility = mean drift velocity v acquired under steady unit field, so that current for unit field is Nev.

$$R = \frac{2m\,(\delta l)^2}{Ne^2\,\tau} \quad .. \quad .. \quad .. \quad (14)$$

Combining equations (13) and (14)

$$< I^2_{df} > R = 4mu^2\,df \quad .. \quad .. \quad .. \quad (15)$$

But if u^2 is equated to the mean-square velocity of an electron in classical thermal equilibrium,

$$mu^2 = kT \quad .. \quad .. \quad .. \quad (16)$$

and this leads to the standard result

$$< I^2_{df} > R^2 = < V^2_{df} > = 4\,R\,kT\,df \quad (17)$$

Now the point is that the fundamental phenomenon is described by equation (13), and it is only by means of (14) and (16) that it can be described in terms of the macroscopic parameters R and T.

The model used above is fallacious (a) because metallic conduction cannot legitimately be described by equation (14) and (b) because equation (16) is not consistent with the modern view that the conduction electrons in a metal have a Fermi-Dirac distribution, but these two deficiencies are self-compensating. It was shown by Lorentz[9] that the conductivity of a metal is proportional to $\int lu \frac{\partial}{\partial u}[f(u)]\;du$ where l is the mean free path of an electron of velocity u and $f(u)$ is the proportion of the total number of electrons which have thermal velocity u; and it has since been shown by Bakker and Heller[10] that there is a universal relationship between conductivity, calculated in this way, and the fluctuation current due to the statistical fluctuations in the distribution of thermal velocities. Hence equation (17) is derivable from the mechanism of metallic conduction, though not so simply as was suggested in equations (12) to (16).

Non-Equilibrium Systems.

It should be noted that there is here no postulate of overall thermodynamic equilibrium between the conductor and external bodies; it is only necessary to know the mean free path and the distribution function of the velocities. In fact Brillouin[6] demonstrated theoretically that the noise would not be altered by the superposition of a drift velocity due to the passage of a steady current of any value for which the resistor remained ohmic in behaviour, and in view of the i^2R heating associated with such a current there is clearly no overall equilibrium in this case. It will, of course, be argued that this is a special case in which the two effects can be separated by virtue of the principle of superposition in a linear system. It is true that the separation is rendered simple by the linearity in the case of a metallic conductor, but in principle they could just as well be separated in other and more difficult cases. Even in the thermionic valve, which has proved so controversial, the following points cannot be disputed:

(a) The apparent temperature of the conductance of a negative grid provided by electron-loading at high frequencies is directly related to cathode temperature.

(b) The 'space-charge-smoothed' shot noise can be expressed in terms of an equivalent temperature of the r_a of the valve which is a fraction of the cathode temperature, though the precise value of the fraction is still uncertain.

(c) The velocity-distribution of an electron stream is not changed by passage through an electrostatic potential barrier in a plane system, but can be changed by using a magnetic field to form a rough beta-ray spectroscope. The resulting change in noise as observed by F. C. Williams[11] was shown by the author to be readily interpretable in terms of the modified distribution of thermal velocity in the magnetically-filtered electron stream.

(d) It has recently been shown by Freeman[12] that, in the use of a temperature-limited diode at frequencies such that the transit time is appreciable, noise due to the thermal velocities of the electrons in transit can be detected.

(e) In the use of gas-discharge tubes as noise-sources at very high frequencies, it is agreed that the 'electron temperatures' are far above the temperatures of the surrounding gas molecules, yet the noise can be adequately described in thermal terms in spite of this lack of equilibrium.

The conclusion is that to calculate the random noise in a circuit one needs to know $\sum e^2\,u^2\,\tau$ where u is the random velocity and τ the time of flight of a charge-carrier. Provided this quantity can be determined in one way or another there is no need to worry about thermodynamic equilibrium.

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NEW BOOKS

Statistics for Technologists

By C. G. PARADINE, M.A.(Cantab.) and B. H. P. RIVETT, M.Sc.(Lond.). Pp. 288 + vii. English Universities Press, Ltd., Warwick Square, London. Price 25s.

Although statistics are usually regarded as a branch of mathematics, statistical methods must necessarily be used by practical workers in many fields. Fully-qualified statisticians are seldom available to direct the application of these methods. The subject is difficult for research workers who are not mathematicians; wrong conclusions can easily be drawn by mishandling data and by poor design of experiments, particularly subjective experiments.

Most books on statistics are either textbooks for the fully-equipped mathematician, or else they are accounts of methods only, in which the underlying theory is largely omitted. 'Statistics for Technologists' achieves a very satisfactory compromise, as might be expected when one author is a Senior Lecturer in Mathematics at Battersea Polytechnic and the other Head of the Field Investigation Group of the National Coal Board. Full details of all numerical methods used are given, with worked examples and explanations of many computational labour-saving devices. A mathematical argument, however, is not avoided when the theory underlying formulae and rules of procedure is explained. Elementary knowledge of algebra and calculus are sufficient for appreciation of the line of argument except when there is a warning asterisk. Sufficient statistical tables are provided and there are some 180 examples, with answers, at the ends of chapters. Reference is made to mathematical textbooks (notably M. G. Kendall's 'Advanced Theory of Statistics') for formal proofs involving advanced mathematics.

The first four chapters deal with frequency distributions in general, statistical parameters, probability, and the binomial, normal and Poisson distributions. The relations between these distributions, and the significance of Sheppard's corrections, moments, skewness, etc., are particularly well explained. Chapter 5 deals with sums of squares of normal variates and χ^2; Chapter 6 with the theory of small samples, variance ratio, Student's t and Fisher's z. Formulae for the probability distributions of χ, t and z are included. Quality control is discussed in Chapter 7, while Chapter 8 deals with single, double and sequential sampling inspection schemes. In Chapter 9 the theory of errors is discussed, including combinations of errors that are normally or rectangularly distributed. Chapter 10 is concerned with the method of least squares, orthogonal polynomials, Fourier series, and approximation to a continuous function. Chapter 11 deals with regression lines and correlation for two and for three variates; rank correlation is briefly mentioned. The analysis of variance (for one factor, two factors, three factors, and two factors with replications) is the subject of Chapter 12, while the principle of maximum likelihood and probit analysis are discussed in Chapter 13.

The book is very clearly printed, and only trivial misprints appear to have occurred, except in the example on page 160. Here the value 16·85 among the repeated observations must be omitted entirely and the variance of the sample is ten times too small. If the value 16·85 was intended for rejection because it is obviously abnormally large, this should have been indicated.

One minor suggestion is that, in any future edition, the Escalator method* should be added to those suggested for solving linear simultaneous equations in section 10.9.2, as this method seems to bring in during the process of calculation quantities useful for least-squares and multiple-correlation problems.

The book can be specially recommended as showing the practical worker where statistical technique is relevant to his work, and how to apply it. The subject is fundamentally mathematical, and this rather frightens many who feel that their mathematical equipment is probably inadequate. Such people should read 'Statistics for Technologists' before giving up hope of being able to master the subject sufficiently for practical purposes.

J. W. H.

* J. Morris. "An Escalator Process for the Solution of Linear Simultaneous Equations". *Phil. Mag.*, Feb. 1946, Vol. 37, pp. 106-120.

Scintillation Counters

By J. B. BIRKS, B.A., Ph.D., F.Inst.P. Pp. 148 + viii. Pergamon Press Ltd., 242 Marylebone Road, London, N.W.1. Price 21s.

Following an introduction which discusses quite briefly the detection of atomic and nuclear radiation and the visual and multiplier types of scintillation counters, there are chapters on the scintillation counter, the photomultiplier tube, pulse height and time resolution. There are then three chapters on inorganic, organic crystalline, and organic plastic and solution phosphors; the final chapter covers applications of scintillation counters.

The material of the book is most clearly presented and few people with a good background of electronics will have any serious difficulty in understanding it. Mathematics is used where necessary but there is no great amount of it and, what there is, is fairly simple. The book will unquestionably be of great help to the newcomer to nuclear physics but it will also be helpful to those in related fields who find the need for some understanding of the methods used.

W. T. C.

Principles of Transistor Circuits

Edited by RICHARD F. SHEA. Pp. 535 + xxx. Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 88s.

This book has been written by nine authors from the Electronics Laboratory of the General Electric Company (U.S.A.), and it deals in a very comprehensive manner with transistors and their circuitry. The first chapter covers semiconductor principles while the second deals with their 'forms, types and characteristics'.

In Chapter 3, the transistor is regarded more from the standpoint of the user—as a 'black box' with three accessible terminals. Equivalent circuits for low-frequency application are deduced. The following chapters cover the basic principles of the amplifier stage, multistage amplifiers using junction transistors, bias stabilization, power amplifiers and d.c. amplifiers.

High-frequency considerations are introduced in Chapter 9 and three chapters cover the characteristics of the transistor itself, the forms of circuit, and circuit design. Video amplifiers are then treated, followed by oscillators, including multivibrators.

Chapters 14 and 15 cover circuit design by duality and circuit analysis by matrix methods, while Chapter 16 covers feedback amplifiers. There is a chapter on transient analysis and one on large-signal operation; the remaining four cover computer circuits, noise, associated semiconductor devices, and small-signal parameter measurement. The appendixes are headed: Matrix Algebra, Definition of Terms and Bibliography.

The book includes an enormous amount of information about transistor circuits and their design and it covers point-contact transistors as well as both *n-p-n* and *p-n-p* junction types. It is invaluable as a reference book but, as a textbook, it suffers from an excess of generality. The beginner would find the going much easier if more simple circuits had been individually treated in the early stages so that he had a more gradual introduction to new ideas. It would also have been easier for him if more attention had been paid to the definitions of terms. For example, as early as page 5, the term 'minority carrier' occurs. There is no indication of its meaning in the text and it does not occur in the 'Definitions of Terms'.

If the reader does not already know the meaning, which is quite probable in such a new and specialized subject, he has to guess. The natural guess is that minority carriers are the mechanism by which a minor part of the current is carried and, therefore, that they are trivial. It is a wrong guess, however, and one that lands the reader in endless difficulties, for the 'minority' actually refers to free electrons and specifies the condition when the material contains mainly bound electrons and places in the structure—holes—which can contain electrons. Conduction is considered to be mainly by the movement of these holes (although it is really by the movement of electrons into the holes to cause an apparent movement of the holes) and to a minor degree only by the movement of free electrons.

The authors are obviously so steeped in the subject that they have overlooked the beginner's difficulties. The result is, unfortunately that, in spite of the apparent simplicity of the opening, the reader does need prior knowledge of transistors before he starts this book.

One particularly valuable equivalent circuit for the transistor is to be found tucked away in the chapter on matrix methods, just where it is likely to be overlooked on account of the unfamiliarity of the mathematics. It is an equivalent circuit for the earthed-emitter transistor and is nothing more than an ordinary triode valve with two resistors—one between grid and cathode, the other between grid and anode. It is the so-called anode follower or see-saw circuit without the series grid resistor.

W. T. C.

Preferred Numbers B.S. 2045 : 1953

Pp. 10. British Standards Institution, 2 Park Street, London, W.1. Price 2s. 6d.

This booklet explains and defines preferred numbers in the Renard series which are derived from geometric series having common ratios of $\sqrt[5]{10}$, $\sqrt[10]{10}$, $\sqrt[20]{10}$, $\sqrt[40]{10}$ or $\sqrt[80]{10}$ and which are known as the R5, R10, etc., series. The series are the ones agreed internationally by the International Organization for Standardization.

It will come as a shock to radio people to find that the preferred numbers of this Standard are not the ones to which they are accustomed in dealing with resistors and capacitors. The R5 range, for instance, is 1, 1·6, 2·5, 4, 6·3, 10, instead of the familiar 1, 1·5, 2·2, 3·3, 4·7, 6·8, 10, based on $\sqrt[6]{10}$.

MEETINGS

I.E.E.

10th February. "Basic Ground-Wave Propagation Characteristics in the 50–800-Mc/s Band", by J. A. Saxton, D.Sc., Ph.D., and "Ground-Wave Field Strength Surveys at 100 and 600 Mc/s", by J. A. Saxton, D.Sc., Ph.D., and B. N. Harden, M.Sc.

11th February. "A Short Modern Review of Fundamental Electromagnetic Theory", by P. Hammond, M.A.

15th February. "The Teaching of the Subject of Insulating Materials", discussion to be opened by Willis Jackson, D.Sc., D.Phil., F.R.S., at 6 o'clock.

22nd February. "Acceptable Standards of Quality in Sound Broadcast Transmission and Reception", discussion to be opened by J. K. Webb, M.Sc.(Eng.), B.Sc.Tech.

4th March. "Submerged Telephone Repeaters for Shallow Water", by R. J. Halsey, B.Sc.(Eng.), and F. C. Wright; "The Netherlands-Denmark Submerged-Repeater System", by A. H. Roche, B.Sc.(Eng.), and F. O. Roe, B.Eng.; and "British Post Office Standard Submerged-Repeater System for Shallow-Water Telephone Cables, with special mention of the England-Netherlands System", by D. C. Walker, B.Sc.(Eng.), and J. F. P. Thomas, B.Sc.(Eng.).

These meetings will be held at the Institution of Electrical Engineers, Savoy Place, Victoria Embankment, London, W.C.2, and will commence at 5.30 except where otherwise stated.

BRIT.I.R.E.

17th February. "Electronics in Film Making", by W. D. Kemp and B. R. Greenhead, to be held at London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1, at 6.30.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Values for December 1953

Date 1953 Dec.	Frequency deviation from nominal: parts in 10^8		Lead of MSF impulses on GBR 1000 G.M.T. time signal in milliseconds
	MSF 60 kc/s 1429-1530 G.M.T.	Droitwich 200 kc/s 1030 G.M.T.	
1	— 1·3	+ 2	+ 33·1
2	— 1·3	+ 2	+ 31·4
3	— 1·3	+ 3	+ 30·4
4	— 1·5	+ 2	+ 28·6
5	— 1·4	+ 2	N.M.
6	— 1·4	+ 2	N.M.
7	N.M.	+ 2	+ 24·9
8	— 1·3	+ 1	+ 22·6
9	— 1·3	+ 2	+ 20·7
10	— 1·3	+ 2	+ 18·7
11	— 1·3	+ 2	+ 16·4
12	— 1·4	+ 3	N.M.
13	— 1·4	+ 3	N.M.
14	— 1·4	+ 2	+ 10·7
15	— 1·4	+ 3	+ 9·2
16	— 1·4	+ 3	+ 6·3
17	— 1·4	+ 3	+ 5·1
18	— 1·3	+ 3	+ 6·3
19	N.M.	+ 4	N.M.
20	N.M.	+ 3	N.M.
21	— 1·5	+ 3	+ 2·1
22	— 1·5	+ 3	— 1·3
23	— 1·5	+ 3	— 3·8
24	— 1·5	+ 4	N.M.
25	N.M.	+ 4	N.M.
26	— 1·4	+ 3	N.M.
27	— 1·4	+ 4	— 10·8
28	— 1·3	+ 4	N.M.
29	— 1·4	+ 4	— 14·5
30	— 1·4	+ 4	— 14·4
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The values are based on astronomical data available on 1st January 1954. The transmitter employed for the 60-kc/s signal is sometimes required for another service.

N.M. = Not Measured.

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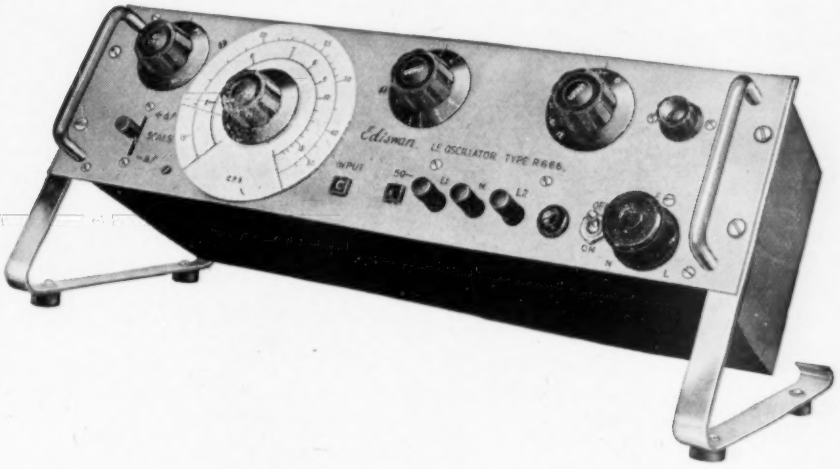

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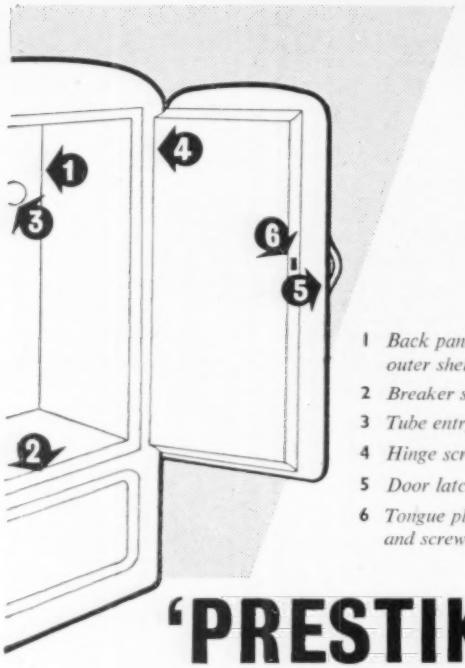

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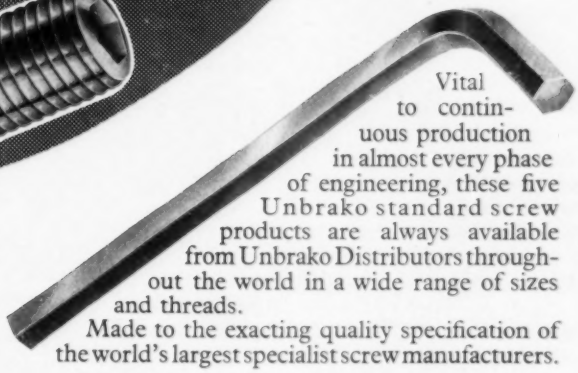

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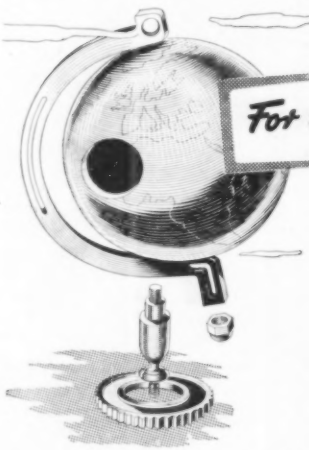

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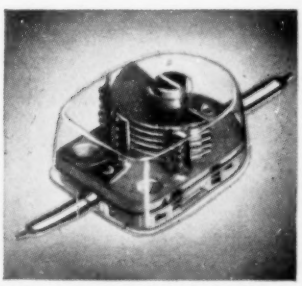

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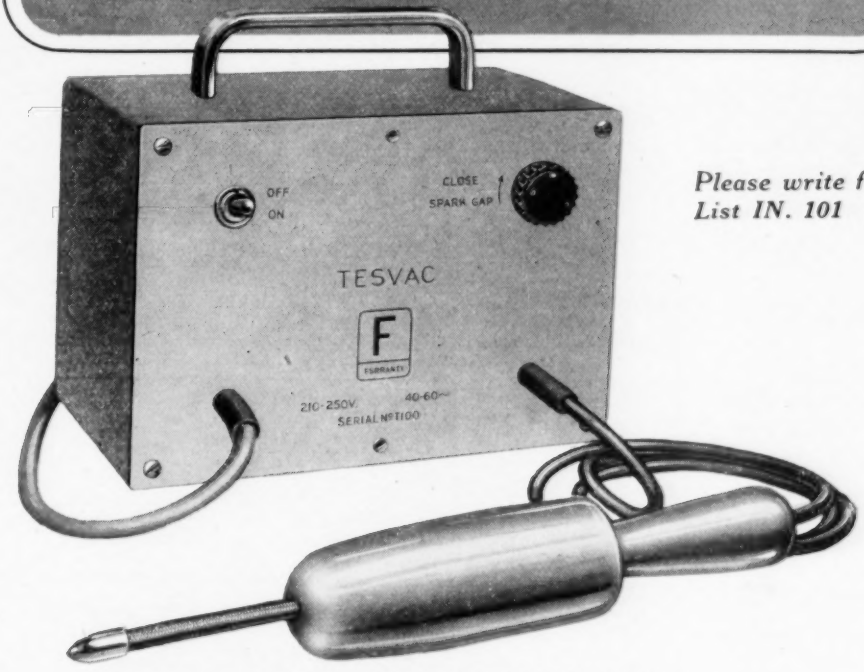

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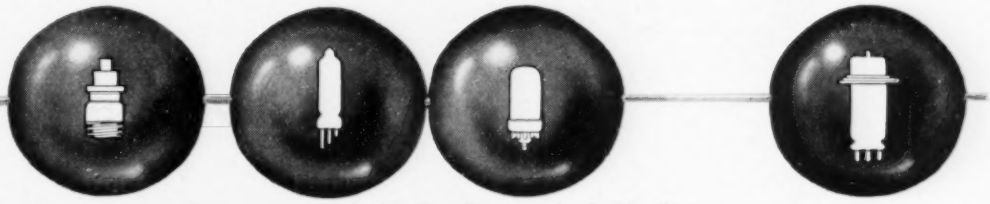

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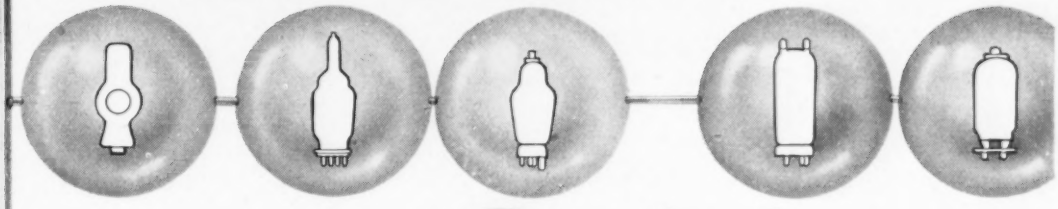

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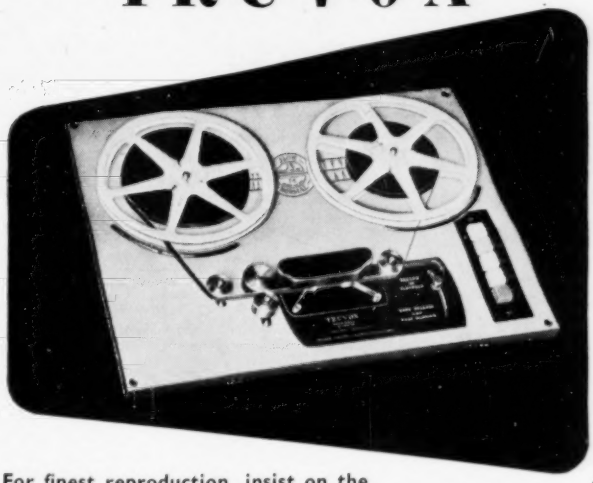

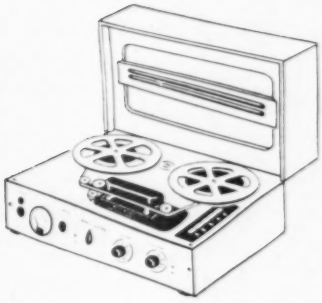

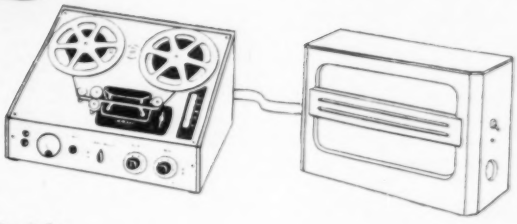

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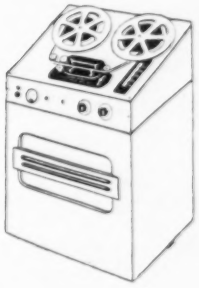

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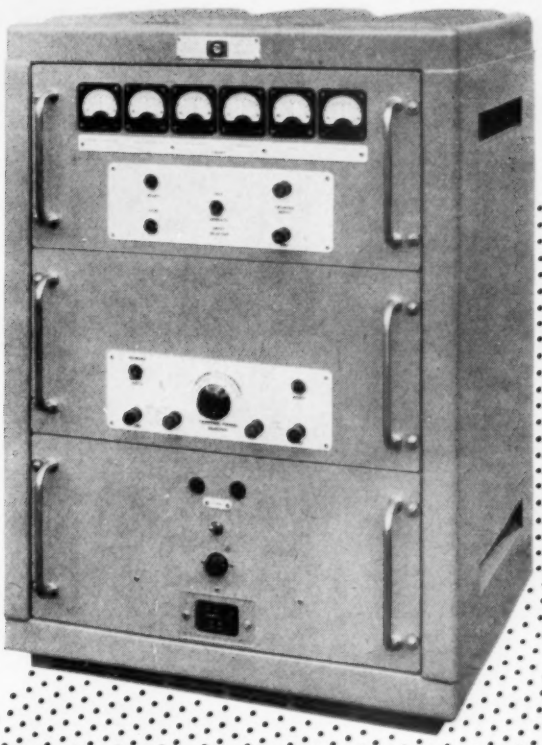

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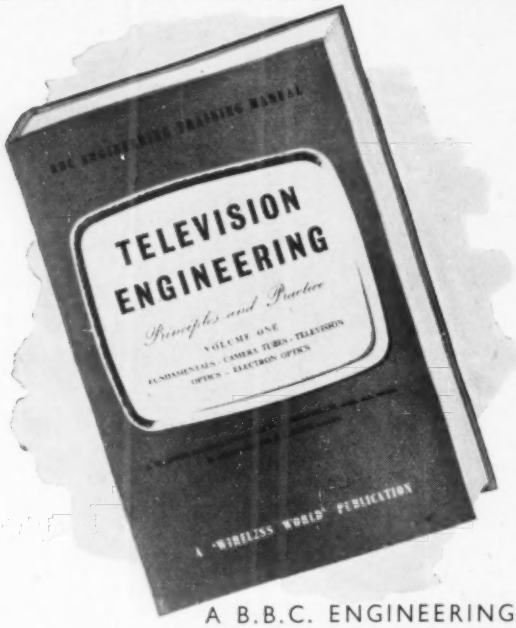

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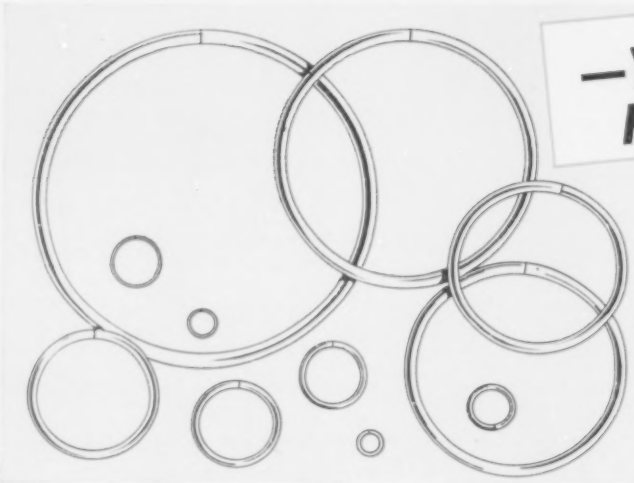

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